Electronic technology

principal contents

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OCTOBER 1961 VOLUME 38 NUMBER 10

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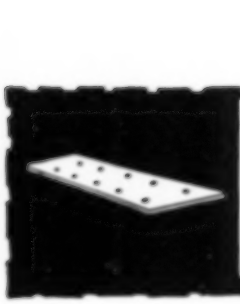

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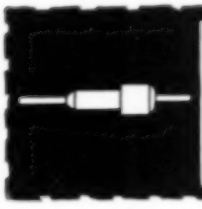

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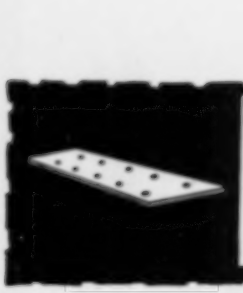

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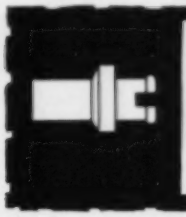

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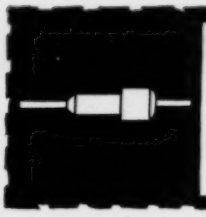

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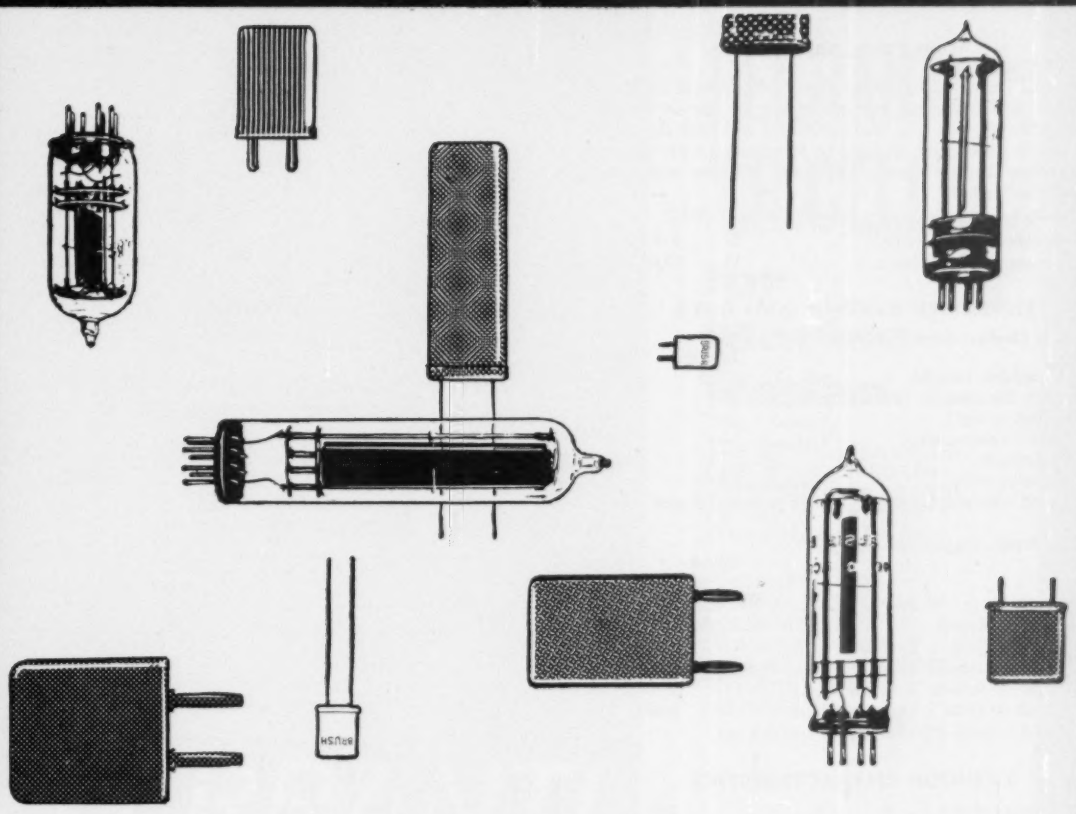

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Heater voltage (volts)	V_h	10.0

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		Triode	Tetrode
Anode Dissipation (watts)	$p_{a(max)}$	1.5	1.5
Screen Grid Dissipation (watts)	$p_{g2(max)}$	—	0.5
Anode Voltage (volts)	$V_{a(max)}$	250	250
Screen Grid Voltage (volts)	$V_{g2(max)}$	—	250
Heater to Cathode Voltage (volts rms)	$V_{h-k(max)rms}$	150*	150*

*Measured with respect to the higher potential heater pin.

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		Triode	Tetrode
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Control Grid to Anode	C_{g-a}	2.4	0.04
Grid Triode to Grid 1 Tetrode	C_{g1-g1}	0.003	
Anode Triode to Anode Tetrode	C_{a1-a}	0.012	
Grid Triode to Anode Tetrode	C_{g1-a}	0.004	
Anode Triode to Grid 1 Tetrode	C_{a1-g1}	0.008	

†Measured in fully shielded socket without can.

TETRODE CHARACTERISTICS

Supply Voltage (volts)	V_b	200	
Anode Load Resistance (kΩ)	R_a	4.7	
Screen Grid Voltage (volts)	V_{g2}	—	80
Control Grid Voltage (volts)	V_{g1}	−0.6	−1.5
Anode Current (mA)	I_a	4	0.1

TRIODE CHARACTERISTICS

Anode Voltage (volts)	V_a	200
Anode Current (mA)	I_a	10
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Amplification Factor	μ	18

Mounting position: Unrestricted.
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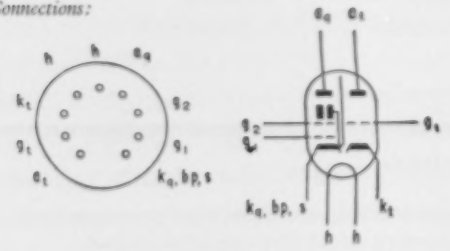

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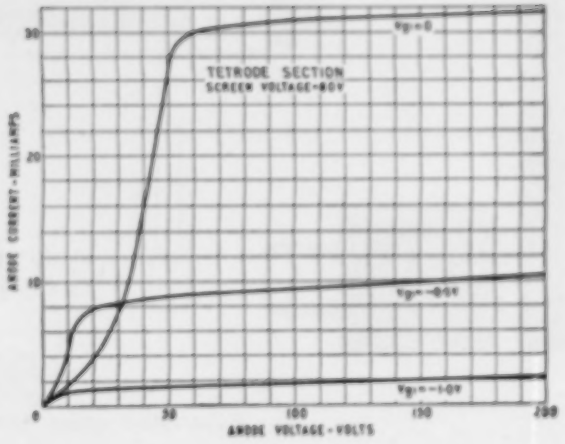

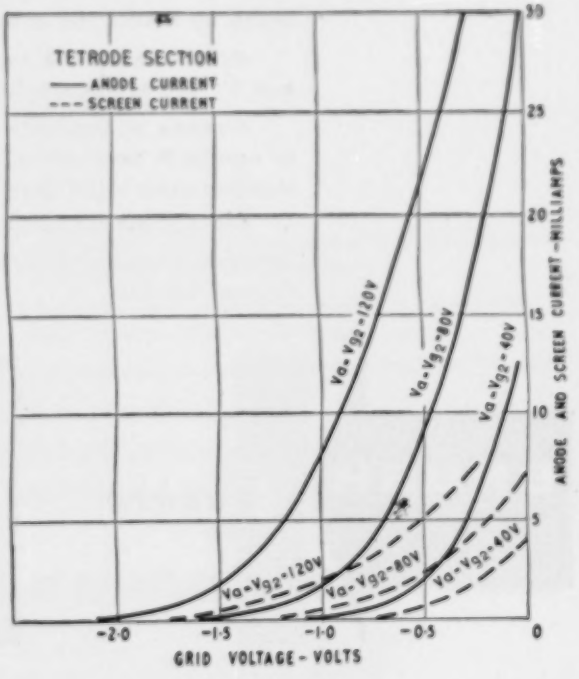

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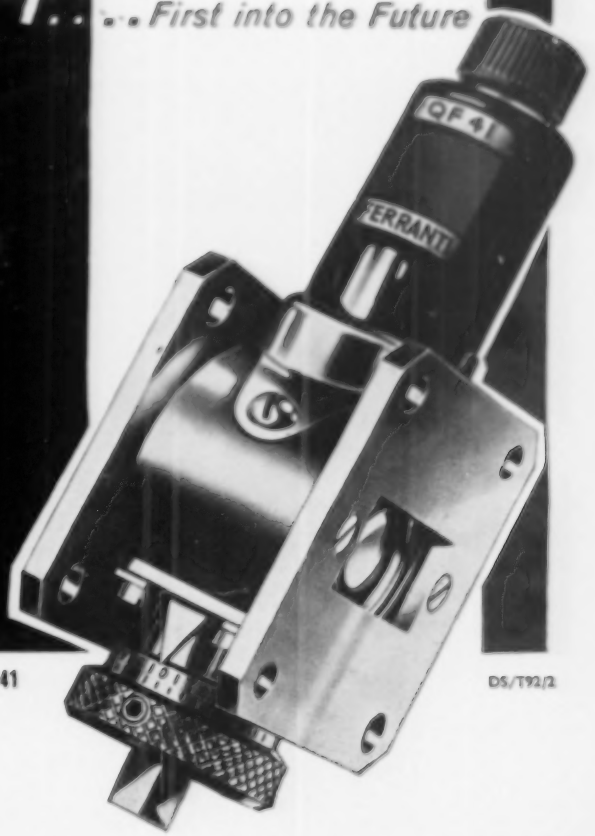

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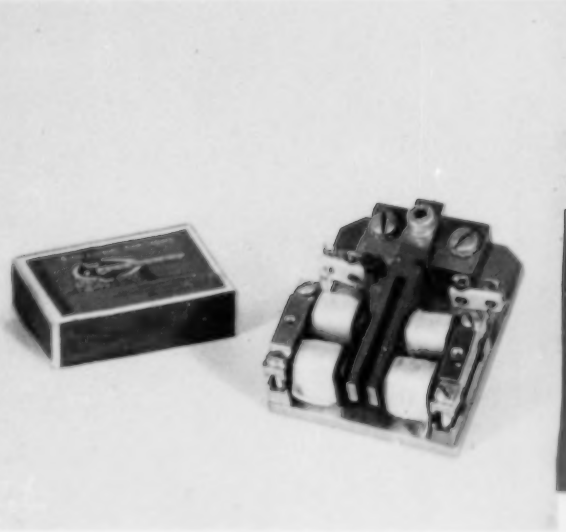

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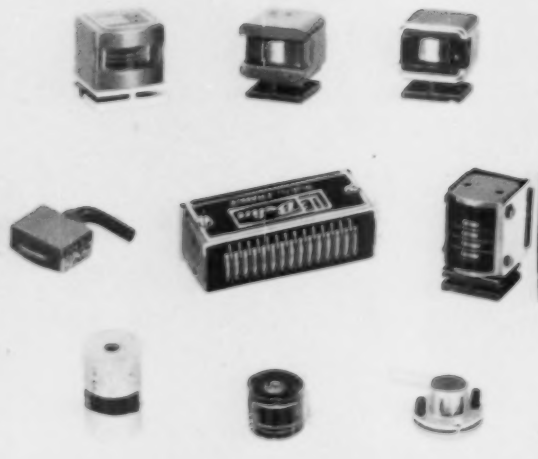

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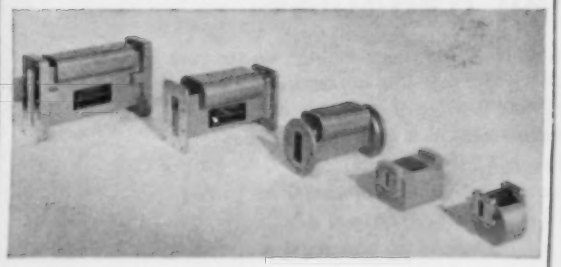

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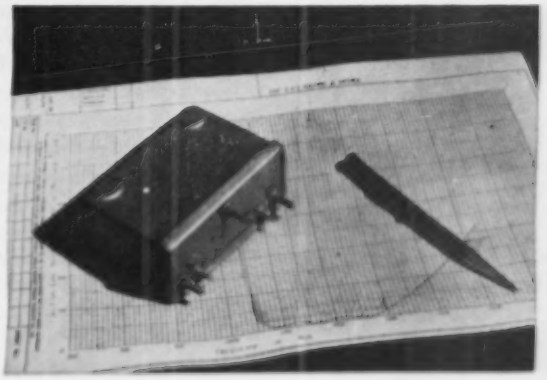

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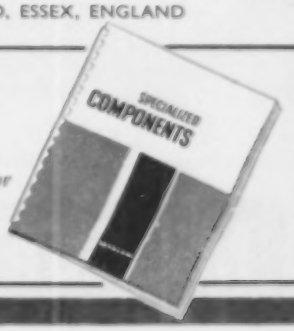

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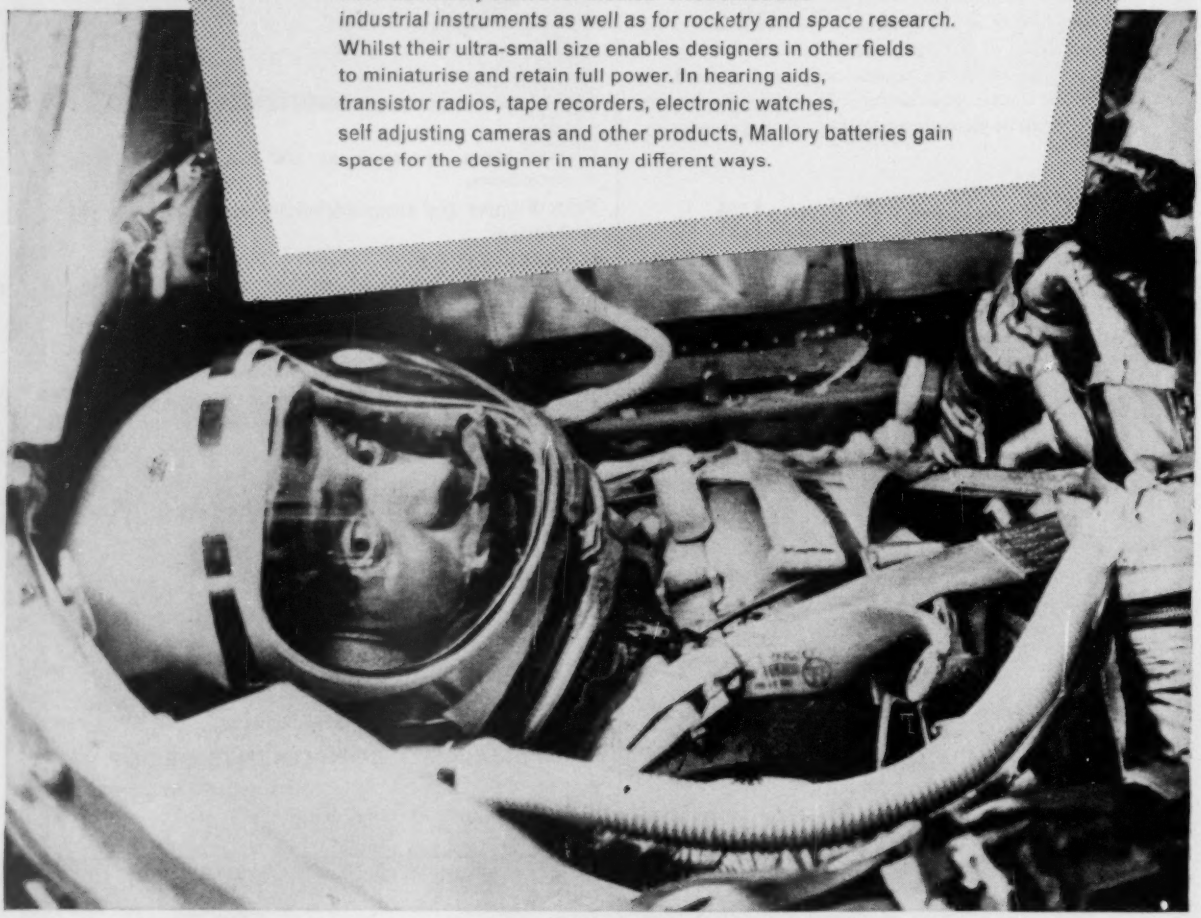

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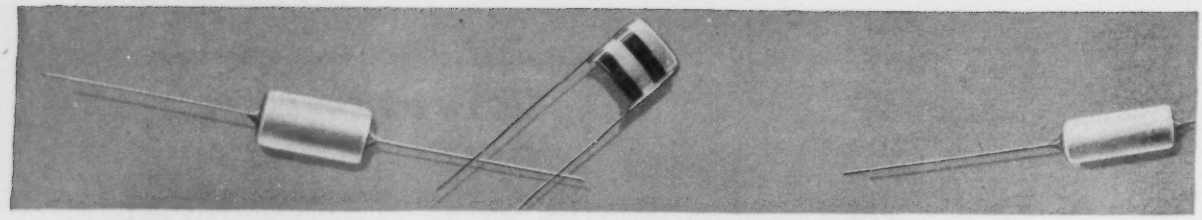

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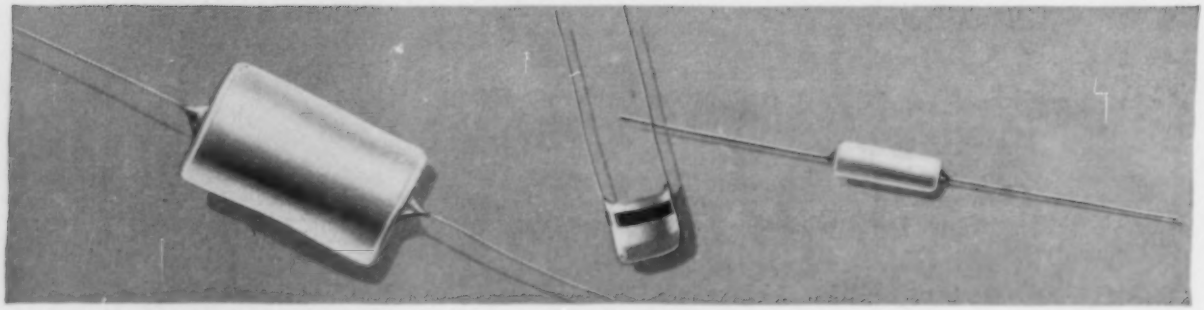

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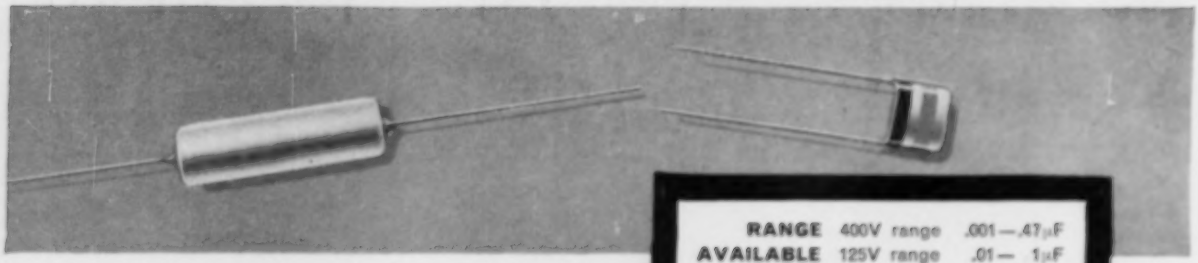

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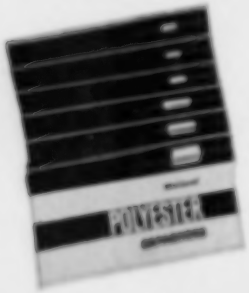

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VOLUME 38 NUMBER 10

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Television Standards

AT this year's National Radio Exhibition the attendance was a good deal larger than last year and it seems likely that it was caused by the demonstrations of 625-line television and of 405-line colour television. For a long time there has been a great deal of talk about both, and these demonstrations were probably the first opportunity that most people had to see for themselves what it is all about.

At the moment colour is not really a live issue for it is generally agreed that it should not be introduced until we have decided what we are going to do about line standards. The decision will not, of course, be made until after the Pilkington Committee has reported next year, and if the eventual decision is for a change of standards it will take quite a long time to implement. Colour television as a service is thus necessarily still quite a long way off.

This is rather a pity, for we suspect that the public is much more interested in colour than in line standards and would rather have colour on 405 lines in the near future than wait until a change of standards has been carried out; if, of course, it is going to be made at all.

Quite a number of people seem to regard it as a foregone conclusion that there will be a change from 405 to 625 lines in spite of the fact that the improvement in picture quality will be quite small and the cost of the change will be very large.

What really does surprise us about this 'lines' controversy is that all discussion is about the number of scanning lines and only rarely are any other television standards mentioned. If we do change to 625 lines, shall we also change to negative modulation for the vision transmission, and shall we also adopt frequency modulation for the sound signal? In other words, do the advocates of 625 lines intend that we change to the Continental 625-line standards in their entirety or do we only change to 625 lines and do as we please in other matters?

No one has yet made clear what is really proposed, and there is a difficulty about adopting Continental standards, for there are two; Eastern Europe has a video bandwidth and a sound-and-vision carrier spacing each 1 Mc/s greater than Western Europe. The Television Advisory Committee recommended that if we change to 625 lines we should use a video bandwidth half-way between the two Continental standards. This appears to indicate that the Committee is not wedded to Continental standards in their entirety!

We really do think that some decision should be made quickly about what 625-line standards we shall adopt if we do change to 625 lines. Receiver manufacturers have to plan a long way ahead and they cannot do very much if the standards are vague. The big 'if' of the change must remain but everything else could be settled.

Voltage Breakdown of Transistors

By S. C. Ryder-Smith, B.Sc.(Eng.), Graduate I.E.E.*

When a p-n junction has a reverse biasing voltage applied to it, only a small leakage current will flow. However, if the applied voltage is raised a point is reached where current flow begins to increase rapidly, with a large increase in current for a small voltage step. When this happens, the junction is said to have broken down. In a simple diode, breakdown may be due either to avalanche or Zener breakdown of the junction. In a transistor, doping levels are usually not high enough for Zener breakdown to occur, but two different mechanisms can cause the transistor to break down. The first is an emitter-multiplied avalanche breakdown; the second is punch-through.

Transistor-voltage ratings are designed to guard against voltage breakdown. They are given in the form of ratings between two terminals, with the third left open-circuit. In addition, derating curves are sometimes given. These show how the ratings may be modified when the third terminal is not left open-circuit, but is connected into the circuit through a finite resistance.

This article describes the breakdown mechanisms in transistors, and deals with some of the cases where external circuit elements are included. Practical circuits rarely conform to the simple conditions for which voltage ratings are given, therefore the question of which rating to apply is discussed.

Breakdown Mechanisms

As stated, breakdown in transistors can be due either to an avalanche effect or to punch-through. In addition, breakdown of the transistor can occur if it is allowed to overheat, in which case permanent damage to the transistor is likely to occur. Voltage breakdown does not, of itself, damage the transistor. However, the rise in dissipation associated with the breakdown may well cause permanent damage through overheating. As breakdown can be a local phenomenon, and some parts of the junction may therefore be hotter than other parts, it is not always safe to assume the same power rating that applies under normal operating conditions.

Avalanche Breakdown

The phenomenon of avalanche breakdown [1,2,3] is discussed here with the case of the simple diode being considered first. Carriers entering the depletion layer of a reverse-biased junction are accelerated across by the field set up by the applied voltage. As the carriers move across, some of their energy will be lost by collisions to the fixed atoms of the crystal lattice. If sufficient energy is transferred in a collision, the fixed atom can become ionized, producing a hole-electron pair. The new carriers thus produced will be accelerated in opposite directions out of the depletion layer, and will therefore contribute to the current flow. When the applied voltage is low, collisions of this sort will be rare and the current flow will be only slightly increased. As the applied voltage increases, so the probability of ionizing collisions rises. A voltage will eventually be reached where the carriers produced in a collision will themselves acquire enough energy to ionize new atoms. When this happens, the current will rise very rapidly with voltage; the voltage at which it happens is the avalanche breakdown voltage of the junction. Fig. 1 shows a typical avalanche breakdown curve.

In a transistor, the collector-base and emitter-base breakdowns (with the third terminal open-circuit) are both due to avalanche multiplication. However, the breakdown between collector and emitter with base open-circuit can be due either to an emitter-accelerated avalanche or to punch-through. When avalanche breakdown occurs, it is at a lower voltage than the breakdown between collector and base, owing to the action of the emitter.

In order best to understand the mechanism of the emitter action, it is perhaps easiest to consider first the difference between the collector-base leakage current (I_{CB0}) and the collector-emitter leakage current (I_{CE0}).

In the case of the collector-base leakage current, the minority carriers leaving the base region and entering the collector depletion layer disturb the balance of charge in the base region. This is restored by the majority carriers leaving the base region through the base contact. When the connection to the base region is made through the emitter (i.e., the base terminal is open-circuit) exactly the same condition holds; charge

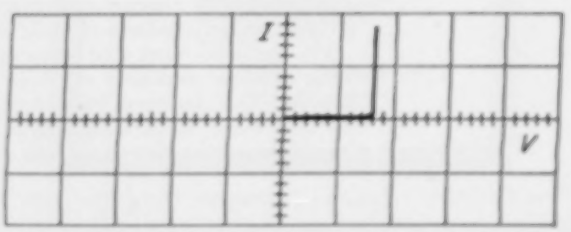


Fig. 1. *Avalanche breakdown in a diode*

* *Standard Telephones and Cables Limited*

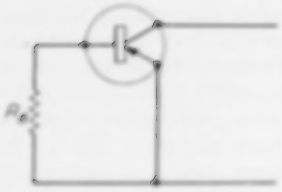


Fig. 2. An external resistance between base and emitter can have an effect on the breakdown characteristic of a transistor

must be supplied to the base at the same rate that it is removed due to the flow of leakage current into the collector. However, only a small fraction of the minority carriers entering the base region are lost in the base by recombination; most reach the collector and increase the total current flow. Hence I_{CB0} is larger than I_{CB0}; in fact it can be shown to be larger by a factor equal to the current gain h_{FB} of the transistor.

When multiplication occurs in the collector, majority carriers generated by multiplication require neutralization in the base region, in the same way as the leakage current. In order to provide this neutralization a much larger emitter current must flow. The collector current will therefore be further increased, and more carriers will be generated by multiplication.

The only difference between the leakage current and the multiplication process is that, in the former, minority carriers leave the base region at the collector; in the latter, majority carriers enter the base region at the collector. At low-current levels it is unlikely that the two processes will result in any difference in the external behaviour of the device.

The result is that when an emitter current flows, avalanche breakdown current is multiplied, and breakdown occurs at a lower voltage than it does with zero emitter current.

A brief mathematical treatment of this effect is given in the appendix, where it is shown that breakdown occurs when collector multiplication counteracts the carrier losses in the base due to recombination. At this point the common-base short-circuit current gain of the transistor is unity, and common-emitter current gain h_{FE} is infinite. This permits a large collector current to flow for zero base current.

Punch-Through

If a reverse bias is applied between collector and base, then the width of the collector depletion layer is usually proportional to the square or cube root of this applied voltage. Thus, it could happen that a voltage lower than the avalanche breakdown could bring the depletion layer across the base until it reached the emitter junction. When this happens, the depletion layer provides an effective short-circuit between collector and emitter, and a large current can flow. This form of breakdown is termed 'punch-through'. The punch-through voltage (V_{PT}) depends on the base width and resistivity. Many transistors are designed to make V_{PT} lie in the region of the collector-base avalanche breakdown voltage. However, there will be large variations from transistor to transistor of a given type. In many, V_{PT} will be above the avalanche breakdown voltage, but there might also be a considerable number in which it will be below; in a small percentage it might even be below the emitter-multiplied avalanche breakdown voltage between collector and emitter.

The Effects of External Circuit Elements

Base Resistance

An external resistance between base and emitter is considered first as shown in Fig. 2.

If the collector-emitter breakdown of the transistor with the base open-circuit is due to punch-through, then the inclusion of R_B will have very little effect on the breakdown voltage. However, if V_{PT} is assumed to be greater than the collector-base avalanche voltage, the inclusion of R_B will have a marked effect on the breakdown characteristic.

One extreme case has already been considered: with

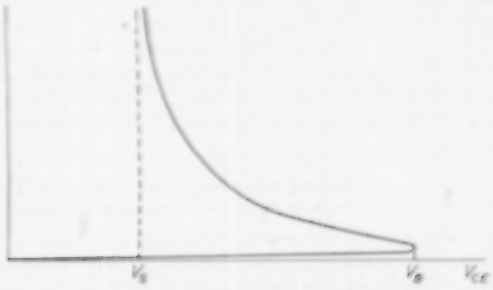


Fig. 3. The breakdown characteristic of a transistor with an external short-circuit between base and emitter. V_S = emitter-multiplied avalanche breakdown voltage between collector and emitter, with base open-circuit; V_B = avalanche breakdown voltage between collector and base, with emitter open-circuit

$R_B = \infty$, it was shown that the normal collector avalanche breakdown was considerably accelerated due to the action of the emitter. The other extreme case, with $R_B = 0$ (although, of course, $r_{bb'}$, the bulk resistance of the base region cannot be eliminated) is now considered.

Before the transistor breaks down, only a very low current (of the order of microamps) flows through the collector; most of this is supplied from the base contact. The effect of the emitter on the avalanche build-up is therefore greatly reduced and the voltage can be raised to the normal collector-base avalanche voltage before breakdown begins. However, once breakdown has started, a larger current flows and this builds up a potential across $r_{bb'}$ which forward-biases the emitter junction. This results in a greater proportion of the collector current flowing out at the emitter than before; the emitter is then able to contribute to the avalanche process. In this way the breakdown voltage relaxes back to the emitter-multiplied avalanche breakdown voltage V_S thus generating a negative resistance region. In avalanche transistors this characteristic is used to give a regenerative switching action, and very fast switching times are possible[4]. A typical curve is shown in Fig. 3.

As the value of R_B is increased, less current is diverted from the emitter before breakdown. In addition, less

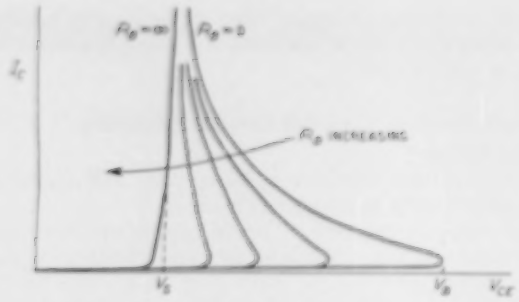


Fig. 4. *Collector-emitter breakdown curves for different values of* R_B

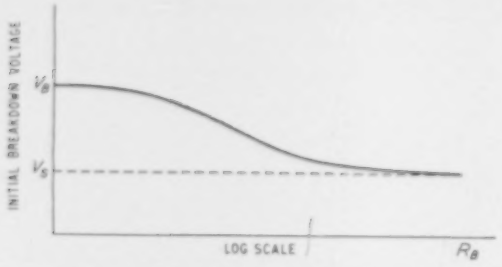


Fig. 5. *Variation of initial breakdown voltage with* R_B

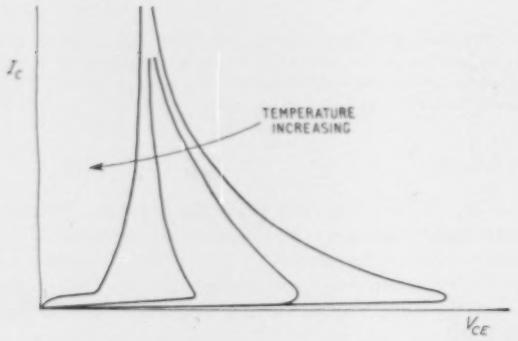


Fig. 6. *Collector-emitter breakdown curves, with* $R_B = 0$, *for different junction temperatures*

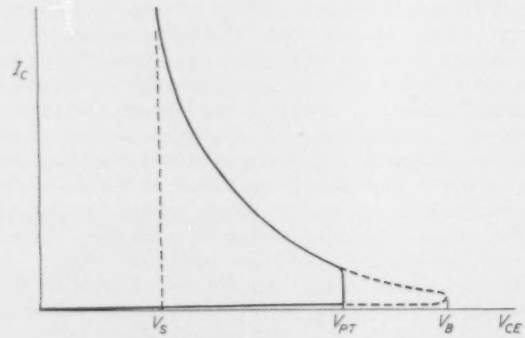


Fig. 7. *Effect of punch-through on the collector-emitter breakdown curve, with* $R_B = 0$

base current is required before the emitter becomes forward-biased sufficiently for it to draw the major part of the collector current. This means that as the value of R_B is increased, the inhibiting effect of the base on the emitter is reduced, and the initial breakdown voltage is also reduced. This is illustrated in Figs 4 and 5. In Fig. 4, a family of breakdown curves is plotted for different values of R_B. Fig. 5 shows the variation of initial breakdown voltage with R_B. The shape of this curve will, of course, depend on how the breakdown voltage is defined. It is usual to define it as occurring at an arbitrary current since this facilitates measurement.

The junction temperature will have a very marked effect on the curves shown in Figs 4 and 5. Breakdown begins when the voltage developed by the leakage current, across $r_{bb'}$ in series with R_B, is sufficient to overcome the emitter junction potential v_t. As the junction temperature rises, I_{CB0}, and hence the voltage across $r_{bb'}$ and R_B increases, and the value of v_t falls. This means that the inhibiting effect of the base on the emitter is reduced at elevated temperatures, and breakdown will occur at a lower voltage. This is illustrated in Fig. 6.

Fig. 6 also shows that raising the junction temperature has an effect similar to that of increasing R_B, except that the low-voltage leakage current is also increased. An increase in junction temperature can be roughly counteracted by reducing R_B where this is possible. This means that the region of transition in the curve in Fig. 5 moves towards the voltage axis as the junction temperature is raised, and a temperature can be reached at which Fig. 5 will appear as a horizontal line at V_S.

It should be emphasized that these curves do not apply if punch-through occurs at a voltage less than V_B. If V_{PT} is less than V_S then breakdown will be due entirely to punch-through. The value of R_B will then have little effect on the breakdown voltage although it might affect the sharpness of the breakdown. When V_{PT} lies between V_B and V_S, a breakdown curve such as is shown in Fig. 7 would result. This curve is for a transistor with an external short between base and emitter.

A mathematical analysis of the behaviour of this circuit is given in the appendix and a brief discussion of its validity is included.

The conclusion must be that a curve of the form shown in Fig. 5 is limited in its application and does not protect the transistor when operated at high currents or high temperature.

Emitter Resistance

In the circuit shown in Fig. 8 R_E provides d.c. negative feedback between emitter and base. This means that the leakage current flowing through R_B is usually not sufficient to initiate an emitter-multiplied breakdown, unless extreme values of R_E or R_B are used. Breakdown, then, will normally begin when the collector-base avalanche voltage (V_B) is reached and a large base current flows. In addition, the voltage approached asymptotically in the high current region of the breakdown curve increases as R_E increases. This is due to the fact that the presence of R_E effectively reduces the transistor gain, and the extent of this effect can in fact

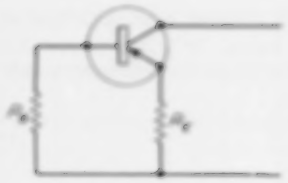
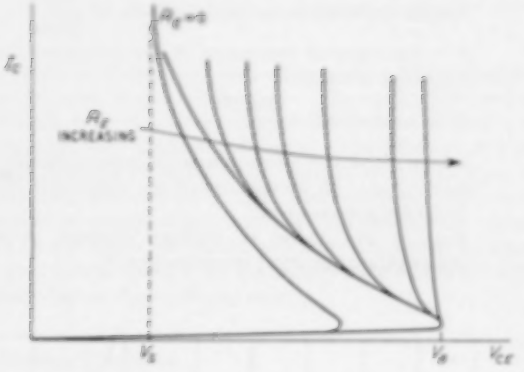


Fig. 8 (*Left*). *External base and emitter resistances can also have an effect on the breakdown characteristics of a transistor*

Fig. 9 (*Below*). *Effect of varying R_E with a fixed R_B*



be specified in terms of the stability factor S of the circuit. Fig. 9 illustrates the effect of varying R_E on the breakdown curve.

If a good stability factor is used, the initial breakdown will not occur until V_B is reached, provided that the punch-through voltage rating is not exceeded. Under these conditions and when the collector current is suitably limited, it is therefore possible to up-rate the transistor to V_B. However, where the collector current is not sufficiently limited, it is of interest to know how the breakdown voltage in the high current region of the breakdown curve varies with the stability factor. A typical derating curve of this sort is shown in Fig. 10.

It is important to note that, in general, derating of this nature cannot be applied where an emitter decoupling capacitor is used and the V_{CE} rating should be used.

Emitter Capacitance

Fig. 11 shows the circuit when an external emitter decoupling capacitance is used. The emitter capacitance C_E can have two effects: one on the breakdown voltage and the other on the circuit behaviour after breakdown. If the voltage across the circuit rises rapidly, the C_E can effectively by-pass R_E and the initial breakdown voltage will be lowered. Once the transistor has broken down, the circuit can start oscillating in a relaxation mode. Fig. 12 shows an oscillogram taken on a curve tracer which illustrates this.

Thus, with an emitter decoupling capacitance, the breakdown behaviour of the circuit is considerably modified and the breakdown voltage becomes more dependent on the rate at which the applied voltage varies. The up-rating of the transistor voltage should therefore be generally confined to the case where the emitter resistance is unbypassed. This is discussed more fully in the section dealing with applications.

Inductive Circuit Elements

When the current through an inductive load is switched off, the voltage across the transistor may be many times larger than the supply voltage. Thus, although a safe value might have been chosen for the supply voltage, the transistor might still break down during switching transients. In many circuits this does not matter, and the breakdown characteristics can serve as a useful means of clipping the voltage spike. However, it is important to ensure that the transistor dissipation remains within safe limits as it passes through the breakdown region. In practice, it is very difficult to give this assurance in circuits where the current is not limited by resistance in the external circuit. This point is discussed further in a later section.

Two things will determine the amount of dissipation: the path taken through the breakdown region, and the rate at which it passes through. However, this question cannot be discussed adequately, nor can equations be written down and conditions for safe operation be deduced until the behaviour of the transistor in the breakdown region is more thoroughly understood.

The breakdown path followed is affected, both by the speed with which it is traced, and by the instantaneous temperature of the active area of the transistor.

The Effect of Tracing Speed in the Breakdown Curve

The equations which have been developed to describe breakdown (see appendix) all include expressions containing h_{FB}. This suggests that the breakdown curves will depend on the speed with which they are traced, since h_{FB} is a complex quantity that is frequency-dependent. The situation is complicated because the frequency characteristics of the transistor alter as the applied voltage rises, due to the narrowing of the base region. The h_{FB} cut-off frequency can increase several times as the punch-through voltage is approached.

Figs 13 and 14 show the breakdown characteristics of an S.T.C. transistor type TK23C as displayed on a curve tracer. The tracing voltage waveform used was

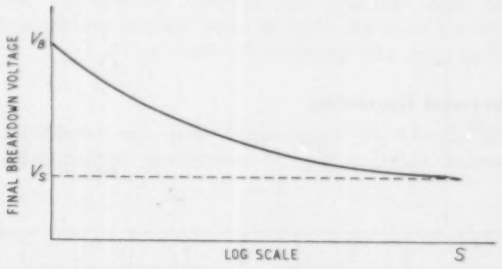


Fig. 10 (*Above*). *Variation of final breakdown voltage with the stability factor S*

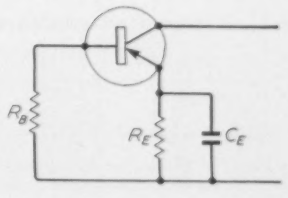


Fig. 11 (*Right*). *Shows the circuit with an external emitter decoupling capacitor*

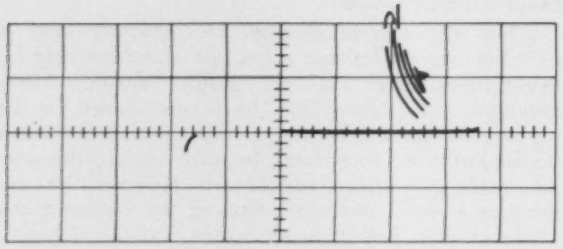


Fig. 12. *Tracing of an oscillogram showing oscillation in the breakdown region of a transistor with an emitter decoupling capacitance*

triangular with approximately equal rise-and-fall times. In Fig. 13 the total tracing time was 2·6 msec and in Fig. 14, 13 μsec.

Figs 15 and 16 are similar to Figs 13 and 14 except that an emitter resistance of 1 kΩ and a base resistance of 18 kΩ have been inserted.

The anti-clockwise paths traced out round the loops of Figs 14 and 16 correspond to an inductive rather than a capacitive reactance, so there is no question of the effect being due to collector capacitance. It can probably be explained in terms of the charge in the base region.

It should be remembered that the path taken in the breakdown region depends on how the current divides between base and emitter. This in turn depends on the total current and on the voltage developed across the base resistance as this forward-biases the emitter junction. In the case of d.c. or low-speed exploration, the current will divide in such a way that the base current forward-biases the emitter sufficiently to allow the remaining current to flow across the emitter junction. However, when fast tracing speeds are used, the diffusion capacitance will have a marked effect on the base potential throughout the tracing cycle. On the rising trace, the diffusion capacitance will inhibit the build-up of voltage across the base resistance, and the transition from V_B and V_S in the breakdown curve will occur at higher current levels. On the falling trace, the diffusion capacitance will tend to hold the base voltage at a high value. This means that a larger proportion of the current will flow through the emitter, particularly at low collector currents, than at slow tracing speeds, and the falling path will therefore lie closer to V_S.

Thermal Runaway

It should be emphasized that the conditions for thermal stability are of overriding importance; the applied voltage might, on this account, have to be limited to a value below the rated value. Whether or not this is the case can be determined from the following condition for thermal stability:

$$V < \frac{1}{S\,\theta\,K\,I_{CB0}} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (1)$$

where

V is the applied voltage

S is the stability factor of the circuit $= \dfrac{\Delta I_C}{\Delta I_{CB0}}$

θ is the thermal resistance of the transistor (in a heat sink if used).

K is a constant $= \dfrac{\log_e 2}{T_n}$

where T_n is the temperature change to double or halve I_{CB0}. (K is typically of the order of 0·07 to 0·09 for germanium).

I_{CB0} is the value of leakage current at the appropriate junction temperature T_j.

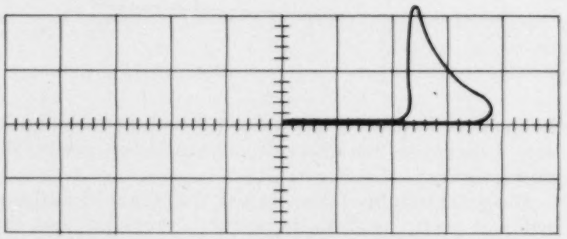


Fig. 14. *As Fig.* 13, *with the tracing time* = 13 μsec

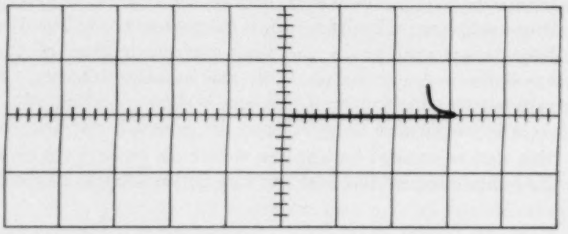


Fig. 15. *As Fig.* 13, *but with* R_B = 18 kΩ *and* R_E = 1 kΩ

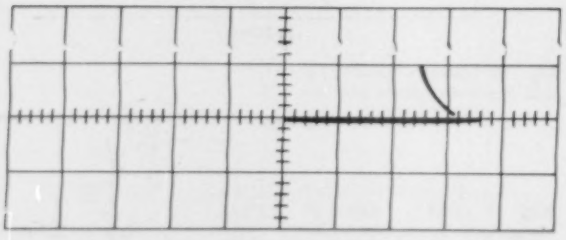


Fig. 13. *Breakdown characteristic of a transistor type TK23C with base and emitter short-circuited externally; tracing time* = 2·6 *msec*

Applying Voltage Ratings to Practical Circuits

The object of a rating should be to protect the transistor against abuse without restricting its use unnecessarily. To achieve this aim with a transistor we are faced with considerable complication; our concern is with the problem of whether or not it is safe to run the transistor at voltages between V_S (V_{CEmax} with base open-circuit) and V_B (V_{CBmax} with emitter open-circuit).

The solution in which the voltage is limited to V_S under all conditions is an extremely simple one, but is unnecessarily restrictive; the alternative is a system of ratings based on the circuit application of the transistor.

Fig. 17 is a diagram from which the transistor's rating in a given application may be derived. The single-hatched area is one in which it is conditionally possible

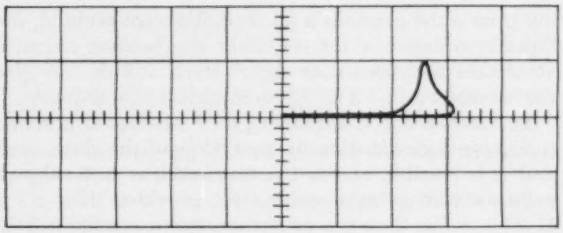


Fig. 16. *As Fig.* 15, *with the tracing time* = 13 μsec

to obtain a stable d.c. operating point. The double-hatched area is not obtainable. The unshaded area is available subject to the normal limitations of current, dissipation and thermal runaway. The abscissa is V_{EB} rather than I_C, though in the positive quadrant V_{EB}, is logarithmically related to I_C. The boundary ABC is a function of temperature, and may vary from one transistor to the next. It would probably be better to treat the problem as though the boundary were ADC, or even slightly below the ordinate axis.

Switching Circuits

We shall consider a transistor having two stable states in its circuit at P and R, and being switched between them. First we shall consider a collector load consisting of a capacitance shunted by a resistance. Starting from P, the switching locus might be P Q_1 R S_1 P. Although this locus crosses the potentially-dangerous region, any instability in that region will tend to aid the switch-on; and, as we have seen in an earlier section, the instantaneous value of V_{CE} is higher during a fast switch-on. The case where the collector load is inductive is shown by the locus (dotted line) P Q_2 R S_2 P. There is a danger that on switch-off the transistor could remain at S_2 for a time sufficient to result in its destruction due to the very high peak dissipation at that point. In view of the very large number of variables involved it is very difficult to establish a criterion of safety for this type of operation. It is generally advisable to limit the operation of circuits with inductive loads to a voltage less than V_S. Under special circumstances the manufacturer may permit exceptions to be made to this rule. A purely resistive load line has a locus P Q_3 R Q_3 P. This is safe provided the base circuit impedance is low enough to ensure that at turn-off a positive voltage appears across the base-emitter terminal under the most adverse conditions of leakage current and temperature. It is good practice to make the dissipation at the point Q_3 less than the continuous maximum dissipation rating of the transistor. The transistor may remain at Q_3 if the positive-bias supply to the base terminal fails. The above discussion is summarized in Fig. 18.

We consider next a circuit in the off-condition where the base is not returned to a positive potential, but is connected through a resistor to earth. The quiescent point P′ lies on, or just above, the ordinate axis in Fig. 17. The results of Fig. 18 still apply, but with some very restrictive limitations. V_{CE} is limited to a value between V_S and V_B given by a derating curve, and this derating curve is a marked function of temperature. Above a junction temperature, typically of the order 45°C to 50°C in a germanium transistor, the value of V_{CE} is limited to V_S. Furthermore, if the circuit is switched from the 'on' to the 'off' state, the switching transient applied to the base terminal must be a positive pulse from a low-impedance source. These are severe limitations, and circuits of this kind must be regarded as being

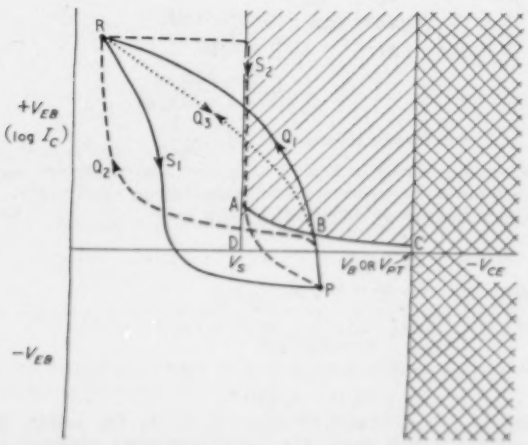


Fig. 17. *Locus diagram with various types of load (p-n-p transistor)*

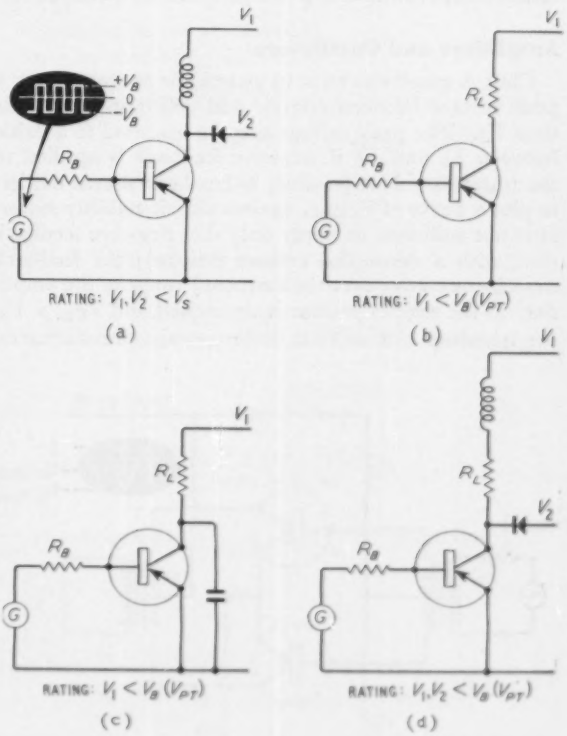


Fig. 18. *Four switching circuits with positive base voltage in the off state. The condition to be met in good design practice for all circuits is that:* $\frac{V_B}{R_B + r_{bb'}} > I_{CB0max}$. *In addition, for circuits (b), (c) and (d) it is necessary to have* $\frac{V_1^2}{4R_L} < P_{Cmax}$

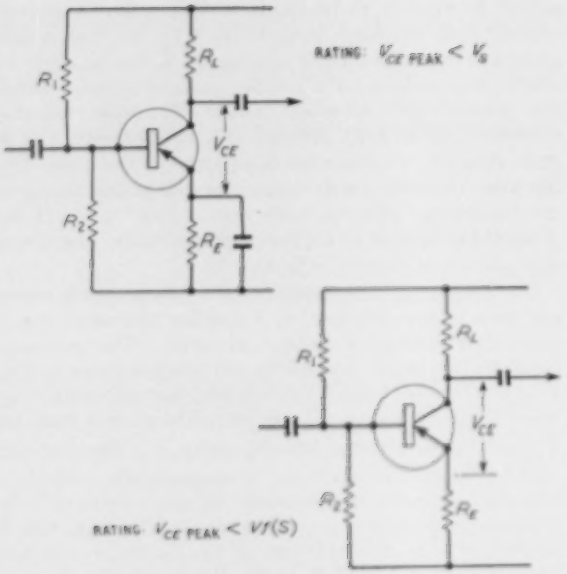


Fig. 19. *Ratings for class A amplifiers [$Vf(S)$ is a voltage that is a function of the stability factor S]*

based on dubious practice. Certainly they are limited to conditions of low ambient temperature and dissipation.

Amplifiers and Oscillators

Class A amplifiers must in general be operated with a peak voltage between emitter and collector that is less than V_S. The peak voltage may be up-rated to a value between V_S and V_B if negative feedback is applied to the transistor. It is possible, as has been shown earlier, to plot a curve of V_{CEmax} against circuit stability factor. It is not sufficient to apply only d.c. negative feedback (i.e., with a decoupled emitter resistor); the feedback must be operative over the frequency range of the amplifier. If the emitter resistor is decoupled and $V_{CE} > V_S$, the transistor can oscillate under certain circumstances;

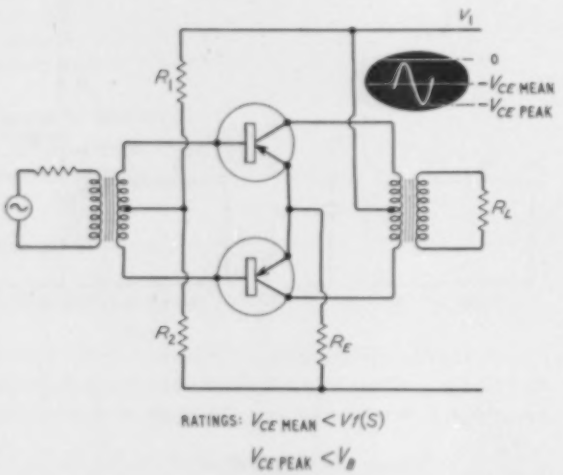


Fig. 20. *Ratings for a class B or C amplifier [$Vf(S)$ is a voltage that is a function of the stability factor S]*

but even if the conditions for oscillation are avoided, the input impedance of the transistor can become negative at certain frequencies or signal levels. This can give rise to distortion. Fig. 19 summarizes the situation.

In classes B and C amplifiers, the base-emitter junction is reverse-biased during at least 45% of the duty cycle and it is possible to run the transistors at a 'no-signal' collector voltage approaching V_S, provided $2V_S < V_B$. If $2V_S > V_B$, then the maximum mean voltage is $\frac{1}{2}V_B$. Assuming $2V_S < V_B$, it is possible to increase the mean voltage above V_S by the use of negative feedback, as in the class A amplifier. In cases where the collector load is inductive and the amplifier is providing a rectangular output pulse, it will be necessary to provide a very low impedance source in the base circuit to turn the transistor off under the most adverse conditions. Some form of clamp to limit the collector voltage to V_B may also be necessary. Fig. 20 summarizes the situation.

In sine-wave oscillators with a transistor or transistors operated in class A, the limitations of the class A amplifier apply. With class B or class C circuits, the limitations of the class B and class C amplifier circuits apply.

The limitations of switching circuits with inductive loads apply in general to square-wave oscillators such as those used in d.c. convertors.

Acknowledgements

The author wishes to express his gratitude to Dr. R. W. A. Scarr for the help which he has given, and to thank Standard Telephones and Cables Limited for permission to publish this article.

APPENDIX

Collector-Emitter Breakdown with Base Open-Circuit

The collector multiplication factor M is defined as the ratio of holes (in the case of a p-n-p transistor) entering the collector from the base to the total collector current[1,2]. At high voltages where multiplication takes place, the emitter-collector current gain h_{FB} is given by

$$h_{FB} = h_{FB0}\,M \qquad .. \quad .. \quad .. \quad .. \quad .. \quad .. \quad (2)$$

where h_{FB0} is the low voltage value of h_{FB}, when $M = 1$.

If the emitter efficiency is assumed to be unity, then the emitter recombination current is $I_E(1 - h_{FB0})$. When the base is open-circuit, this must equal the collector base leakage current (I_{CB0}) plus the current generated by collector multiplication.

$$\therefore I_E(1 - h_{FB0}) = -I_{CB0} - I_C(1 - 1/M)$$

But $$I_C = -I_E$$

$$\therefore I_C(1 - h_{FB0}) = I_{CB0} + I_C(1 - 1/M)$$

$$\therefore I_C(1 - h_{FB0} - 1 + 1/M) = I_{CB0}$$

$$\therefore I_C = \frac{M\,I_{CB0}}{1 - M\,h_{FB0}} \qquad .. \quad .. \quad (3)$$

$$\therefore I_C \text{ will become infinite when } M = \frac{1}{h_{FB0}} \qquad .. \quad .. \quad (4)$$

Substituting for M in Equ. (2) $h_{FB} = 1$, thus breakdown occurs when the emitter-collector current gain becomes equal to one.

As a function of voltage, M is given by

$$M = \frac{1}{1 - (V_C/V_B)^n} \qquad .. \quad .. \quad .. \quad .. \quad .. \quad .. \quad (5)$$

when $V_C = V_S$, $M = 1/h_{FB0}$

$$\therefore \quad \frac{1}{h_{FB0}} = \frac{1}{1 - (V_S/V_B)^n}$$

$$\therefore (V_S/V_B)^n = 1 - h_{FB0}$$

$$\therefore \quad V_S = V_B(1 - h_{FB0})^{1/n} \qquad .. \quad .. \quad .. \quad .. \quad (6)$$

But $(1 - h_{FB0})^{1/n} \simeq \beta_0^{-1/n}$ where β_0 is the low voltage base-collector current gain. For p-n-p germanium transistors, n is approximately 3.

$$\therefore V_S \simeq \frac{V_B}{\beta_s^{1/3}} \quad \text{or} \quad V_B \simeq V_S \beta_s^{1/3} \qquad \ldots \quad \ldots \quad \ldots \quad (6a)$$

Hence, for a transistor with a low voltage gain of 64, the collector-emitter breakdown, with base open-circuit, would occur at a quarter of the collector-base avalanche breakdown voltage.

Collector-Emitter Breakdown with an External Resistance between Base and Emitter

The circuit under consideration is shown in Fig. 21. The relevant equations are:

$$I_C = h_{FBN} M I_E - M |I_{CB0}| \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (7)$$

$$I_E = \frac{|I_{EB0}|}{1 - h_{FBN} h_{FBI}} [\exp (VE/v_t) - 1 + h_{FBN}] \qquad \ldots \quad (8)$$

$$V_E = -(I_C + I_E) R_B \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

$$M = \frac{1}{1 - (V_C/V_B)^n} \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (10)$$

Equs (7) and (8) are derived from the Ebers and Moll equations[4]. Equ. (9) from an application of Ohm's Law, and Equ. (10) is an empirical relationship[3,5].

Terms previously undefined are:

h_{FBN} h_{FBI} — the normal and inverse common-base short-circuit forward-transfer ratios

$|I_{CB0}|$ $|I_{EB0}|$ — the magnitudes of I_{CB0} and I_{EB0}, the collector and emitter junction leakage currents, irrespective of their signs

v_t — the 'built in' emitter junction potential.

From Equ. (7)

$$I_E = -\frac{I_C + M |I_{CB0}|}{h_{FBN} M}$$

Putting this value for I_E in Equ. (9) gives

$$V_E = -\left\{I_C - \frac{I_C + M |I_{CB0}|}{h_{FBN} M}\right\} R_B$$

Substituting this value for V_E in Equ. (8) gives

$$I_E = \frac{|I_{EB0}|}{1 - h_{FBN} h_{FBI}} \left\{\exp\left[-\frac{R_B}{v_t}\left(I_C - \frac{I_C + M |I_{CB0}|}{h_{FBN} M}\right)\right] - 1 + h_{FBN}\right\}$$

Putting this value for I_E in Equ. (7) gives

$$I_C = \frac{-h_{FBN} M |I_{EB0}|}{1 - h_{FBN} h_{FBI}} \left\{\exp\left[-\frac{R_B}{v_t}\left(I_C - \frac{I_C + M |I_{CB0}|}{h_{FBN} M}\right)\right] - 1 + h_{FBN}\right\} - M |I_{CB0}|$$

$$\therefore \frac{I_C (1 - h_{FBN} h_{FBI})}{h_{FBN} M |I_{CB0}|} = -\exp\left\{-\frac{R_B}{v_t}\left(I_C - \frac{I_C + M |I_{CB0}|}{h_{FBN} M}\right)\right\} + 1 - h_{FBN} - \frac{|I_{CB0}|}{h_{FBN} |I_{EB0}|} \times (1 - h_{FBN} h_{FBI})$$

$$\therefore \exp\left\{-\frac{R_B}{v_t}\left(I_C - \frac{I_C + M |I_{CB0}|}{h_{FBN} M}\right)\right\} = 1 - h_{FBN} - \frac{|I_{CB0}|}{h_{FBN} |I_{EB0}|} (1 - h_{FBN} h_{FBI}) - \frac{I_C (1 - h_{FBN} h_{FBI})}{h_{FBN} M |I_{EB0}|}$$

$$\therefore -\frac{R_B}{v_t}\left(I_C - \frac{I_C + M |I_{CB0}|}{h_{FBN} M}\right) = \log_e \left[1 - h_{FBN} - \frac{|I_{CB0}|}{h_{FBN} |I_{EB0}|} (1 - h_{FBN} h_{FBI}) - \frac{I_C (1 - h_{FBN} h_{FBI})}{h_{FBN} M |I_{EB0}|}\right]$$

$$\therefore R_B = \frac{-v_t h_{FBN} M}{I_C h_{FBN} M - I_C - M |I_{CB0}|} \log_e \left[1 - h_{FBN} - \frac{|I_{CB0}|}{h_{FBN} |I_{EB0}|} (1 - h_{FBN} h_{FBI}) - \frac{I_C}{h_{FBN} M |I_{EB0}|} (1 - h_{FBN} h_{FBI})\right]$$

Substituting Equ. (10) gives

$$R_B = \frac{-v_t h_{FBN}}{I_C h_{FBN} - I_C [1 - (V_C/V_B)^n] - |I_{CB0}|} \log_e \left[1 - h_{FBN} - \frac{|I_{CB0}|}{h_{FBN} |I_{EB0}|} (1 - h_{FBN} h_{FBI}) - \frac{I_C}{h_{FBN} |I_{EB0}|} (1 - h_{FBN} h_{FBI}) (1 - V_C/V_B)^n\right]$$

$$\therefore R_B = \frac{v_t h_{FBN}}{I_C (1 - h_{FBN}) + |I_{CB0}| - I_C (V_C/V_B)^n} \log_e \left[I_C \frac{(1 - h_{FBN} h_{FBI})}{h_{FBN} |I_{EB0}|} \left(\frac{V_C}{V_B}\right)^n + 1 - h_{FBN} - (I_C + |I_{CB0}|) \frac{(1 - h_{FBN} h_{FBI})}{h_{FBN} |I_{EB0}|}\right]$$

Let the breakdown voltage be defined as the voltage at which the collector current rises to an arbitrarily defined value, say $-I_A$. Thus, at breakdown $I_C = -I_A$ and

$$R_B = \frac{v_t h_{FBN}}{I_A (V_C/V_B)^n - I_A (1 - h_{FBN}) + |I_{CB0}|} \log_e \left[1 - h_{FBN} + \frac{(1 - h_{FBN} h_{FBI})}{h_{FBN} |I_{EB0}|} (I_A - |I_{CB0}|) - I_A \frac{(1 - h_{FBN} h_{FBI})}{h_{FBN} |I_{CB0}|} \left(\frac{V_C}{V_B}\right)^n\right]$$

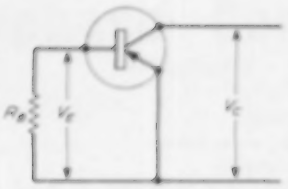


Fig. 21

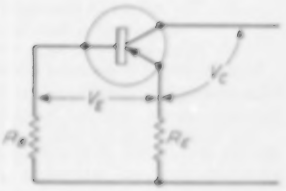


Fig. 22

For simplicity, this may be written

$$R_B = \frac{A}{B (V_C/V_B)^n - C} \log_e [D - E (V_C/V_B)^n] \qquad \ldots \quad \ldots \quad (11)$$

where the constants A, B, C, D and E are given by

$$A = v_t h_{FBN}$$

$$B = I_A$$

$$C = I_A (1 - h_{FBN}) - |I_{CB0}|$$

$$D = \frac{(1 - h_{FBN} h_{FBI})}{h_{FBN} |I_{EB0}|} (I_A - |I_{CB0}|) + 1 - h_{FBN}$$

$$E = \frac{I_A (1 - h_{FBN} h_{FBI})}{h_{FBN} |I_{EB0}|}$$

Once these constants have been evaluated, Equ. (11) enables a value of R_B to be worked out for any given value of the ratio (V_C/V_B).

Considering the extreme cases of R_B

$R_B = \infty$, when $B (V_C/V_B)^n = C$

i.e., $I_A (V_C/V_B)^n = I_A (1 - h_{FBN}) - |I_{CB0}|$

$$\therefore \quad (V_C/V_B)^n = 1 - h_{FBN} - \frac{|I_{CB0}|}{I_A}$$

If I_A is large,

$$(V_C/V_B) = (1 - h_{FBN})^{1/n} \quad \ldots \quad \text{cf. Equ. (6)}$$

$R_B = 0$ when $\log_e [D - E (V_C/V_B)^n] = 0$

i.e., $D - E (V_C/V_B)^n = 1$

This leads to $V_C/V_B = 1$ for I_A large

Causes in inaccuracy in Equ. (11) are as follows:

(a) Surface effects have been ignored.

(b) The effects of transistor heating at breakdown on $|I_{CB0}|$, $|I_{EB0}|$ and v_t have been ignored.

(c) h_{FBN} and h_{FBI} have been assumed to be constant.

In addition, it is doubtful whether the Ebers and Moll equations fully retain their validity when high voltages are applied, and the transistor moves into the breakdown region.

The analysis applies only to low frequency on d.c. exploration of the breakdown curve, since the effects of transistor capacitance on the currents flowing have been neglected.

Collector-Emitter Breakdown with External Base and Emitter Resistances

The circuit is shown in Figure 22.

The analysis given in the above section can also be applied here. Equ. (9) is modified to become

$$V_E = -(I_C + I_E) R_B - I_E R_E \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (12)$$

If (12) is used in the analysis in place of (9), Equ. (11) becomes:

$$R_B = \frac{A}{B (V_C/V_B)^n - C} \log_e [D - E (V_C/V_B)^n] + \frac{F - B (V_C/V_B)^n}{B (V_C/V_B)^n - C} R_E \quad \ldots \quad \ldots \quad (13)$$

where $F = I_A - |I_{CB0}|$

Dependence of the Breakdown Voltage on the Circuit Stability Factor

The breakdown voltage at high current levels is considered, rather than the initial breakdown voltage. The following assumptions are made:

$$I_A \gg |I_{CB0}|$$

$$I_E R_E \gg V_E$$

Equ. (13) can be written

$$R_B [B (V_C/V_B)^n - C] = \underbrace{v_t h_{FBN} \log_e [D - E (V_C/V_B)^n]}_{X} + \underbrace{[F - B (V_C/V_B)^n] R_E}_{Y} \quad \ldots \quad (14)$$

where $X = v_t h_{FBN} \log_e [D - E (V_C/V_B)^n]$

$Y = [F - B (V_C/V_B)^n] R_E$

It is required to show that $Y \gg X$, so that X may be neglected. Now Y is of the order of $I_E R_E$, and X is of the order of V_E multiplied by the log term. Thus, on the basis of the assumption $R_E I_E \gg V_E$, the inequality $Y \gg X$ will hold, unless the log term is very large. For the log term to become large, I_A must become very large indeed and, under these conditions, the log term is of the order of $\log_e (I_A/|I_{EB0}|)$. Therefore, for the log term to have a value of, say, 20 requires that

$$\frac{I_A}{|I_{EB0}|} = 5 \times 10^8$$

$\therefore$ if $|I_{EB0}| = 1\ \mu A \qquad I_A = 500\ A$

It is therefore legitimate to neglect the term X if, to the assumption that $I_A \gg |I_{CB0}|$, is added the condition that I_A is not extremely large (i.e., of the order of many hundreds of amps, a current much too large to be met in practice).

Under these conditions Equ. (14) becomes

$$R_B [B (V_C/V_B)^n - C] = R_E [F - B (V_C/V_B)^n] \quad \ldots \quad \ldots \quad (15)$$

Now, the stability factor S of the circuit is given approximately by

$$S = 1 + R_B/R_E \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (16)$$

Putting Equ. (15) in Equ. (16) gives

$$S = 1 + \frac{F - B (V_C/V_B)^n}{B (V_C/V_B)^n - C}$$

$$\therefore [B (V_C/V_B)^n - C] (S - 1) = F - B (V_C/V_B)^n$$

$$\therefore (V_C/V_B)^n = \frac{F + C (S - 1)}{B S}$$

Substituting values for B, C and F gives

$$(V_C/V_B)^n = \frac{I_A - |I_{CB0}| + [I_A (1 - h_{FBN}) - |I_{CB0}|] (S - 1)}{I_A S}$$

$$= \frac{1 + (1 - h_{FBN}) (S - 1)}{S}$$

$$= 1 - h_{FBN} + \frac{h_{FBN}}{S}$$

$$\therefore V_C = V_B \sqrt[n]{1 - h_{FBN} + \frac{h_{FBN}}{S}} \quad \ldots \quad \ldots \quad \ldots \quad (17)$$

Extreme cases:

$S = 1$ gives $V_C = V_B$

$S \to \infty$ gives $V_C = V_S = V_B (1 - h_{FBN})^{1/n}$ .. cf. Equ. (6)

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COMPUTER OUTPUT PRINTER

A production model of the "Xeronic" printer is being demonstrated in public for the first time at this year's Computer Exhibition. The machine has been designed to print information from computers on to 26-in. wide plain paper at a rate of 4,700 characters per second; simultaneously it prints its own forms to contain the information.

In operation, information to be printed is fed into the machine from a computer or a magnetic tape unit in digital form. After the information is processed, characters are generated and displayed as a pattern of overlapping dots on the face of a c.r.t. and projected on to a rotating selenium-coated drum. In darkness the surface of the drum is positively charged and light falling on to the drum surface from the c.r.t. discharges an area of the drum corresponding to the character on the tube. The charge pattern on the drum is "developed" by cascading over it a positively charged powder which adheres to the discharged areas around the drum.

As the drum rotates it comes into contact with paper fed continuously from a roll and the powdered "printing ink" is transferred from the drum to the paper by means of a negative charge applied to the paper. Thus the characters formed on the c.r.t. are transferred to the paper. They are rendered permanent in a fusing chamber where printing powder is heated and fused to the paper.

The residual charge on the drum is removed by a discharge grid and a discharge lamp and a brush removes any remaining printing powder.

Single-Phase to Polyphase Convertors

AUTOMATIC STATIC DEVICES FOR 50-c/s POWER

By F. Butler, O.B.E., B.Sc., M.I.E.E., M. Brit. I.R.E.

SUMMARY. *In this article static equipment is described for single to two or three-phase 50-c/s energy conversion. Balanced polyphase voltages, for varying symmetrical loads, are maintained automatically by the use of fixed capacitors in conjunction with saturable reactors in which the d.c. excitation and the corresponding reactance are arranged to vary with the power output. Auxiliary manual control of the reactor excitation permits an exact phase-balance at any particular load. The apparatus functions efficiently over a wide range of load impedance and power factor, is not sensitive to small changes of input voltage and the output waveform is of acceptably low distortion. The convertor is reversible.*

The smaller laboratories and workshops are commonly supplied from single-phase a.c. mains and are thereby debarred from installing a wide range of electrical equipment which demands a polyphase supply. Machine tools powered by built-in three-phase motors cannot usually be modified for single-phase drive, either because of the physical size or because of the unacceptable torque fluctuations of single-phase motors. Polyphase rectifier sets are desirable in cases where only a small ripple voltage can be tolerated on the d.c. output or where freedom from interference with associated equipment is essential. Hum modulation of transmitters using valves with directly-heated filaments can be minimized by polyphase operation and there are special control and regulation problems which are most easily solved if a three-phase supply is available. Elementary phase-splitting circuits can be devised to supply small and constant polyphase loads, although difficulties are encountered when the load is variable. To meet this case, equipment has been produced commercially[1], using a combination of capacitors and iron-cored a.c. saturable reactors. The performance and principle of operation of this apparatus have been described in a published paper[2].

The new circuit to be described is a phase-shifting network in which the magnitude of the impedance in the two arms is a function of the output load current. The desired impedance change, with load variation, is accomplished by automatically-variable d.c. excitation of two iron-cored reactors.

Theory of Phase-Conversion Circuit

The most elementary circuit for single to three-phase conversion is shown in Fig. 1.

Referring to Fig. 1, it is clear that the resistor R which is connected directly across the supply mains can have no effect on the relative phases of E_L and E_C. Ignoring it, there remains a network of impedances, the value of which may be written down by inspection, and

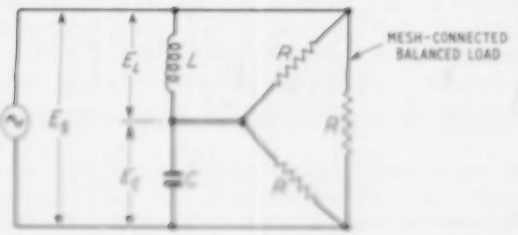

Fig. 1. Elementary phase-conversion circuit. (E_S = r.m.s. supply voltage, E_L = voltage drop across the inductor L, E_C = voltage drop across the capacitor C, R = each element of the mesh-connected pure resistance load.)

from which the desired voltage ratios may be calculated. The conditions will now be established under which the voltages E_L, E_C and E_S are equal in magnitude and symmetrically phased at 120-degree intervals.

Using the component values shown in Fig. 1:

$$E_L : E_C : E_S = \frac{j\omega L R}{R + j\omega L} : \frac{R/j\omega C}{R + 1/j\omega C} : \frac{j\omega L R}{R + j\omega L} + \frac{R/j\omega C}{R + 1/j\omega C} \quad \ldots \quad (1)$$

$$\text{Then } \frac{E_L}{E_C} = \frac{j\omega L R}{R + j\omega L} \cdot \frac{R + 1/j\omega C}{R/j\omega C} = -\frac{\omega^2 LC\{R^2 - L/C - jR(\omega L + 1/\omega C)\}}{R^2 + \omega^2 L^2} \quad \ldots \quad (2)$$

In Equ. (2) set $\omega L = 1/\omega C = R/\sqrt{3}$, so that $L/C = R^2/3$.

$$\text{Then } \frac{E_L}{E_C} = -\frac{1}{2} + j\frac{\sqrt{3}}{2} \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

The two voltages E_L and E_C are thus equal in magnitude and differ in phase by 120°.

NOTE: *All rights on the equipment described have been assigned to the Westinghouse Brake and Signal Company, Limited.*

In the same way, from Equ. (1)

$$\frac{E_S}{E_C} = \frac{\dfrac{j\omega L R}{R + j\omega L} + \dfrac{R/j\omega C}{R + 1/j\omega C}}{\dfrac{R/j\omega C}{R + 1/j\omega C}} = 1 + \frac{\dfrac{j\omega L R}{R + j\omega L}}{\dfrac{R/j\omega C}{R + 1/j\omega C}}$$

$$= 1 + \frac{E_L}{E_C}$$

Hence $$\frac{E_S}{E_C} = \frac{1}{2} + j\frac{\sqrt{3}}{2} \quad .. \quad .. \quad .. \quad .. \quad (4)$$

From Equs (3) and (4), with appropriate sign conventions, it may be seen that E_L, E_C and E_S form a balanced 3-phase system, always subject to the condition that $\omega L = 1/\omega C = R/\sqrt{3}$.

By setting the phase difference equal to 90°, equations giving the conditions for single to two-phase conversion may be derived in a similar manner.

Again referring to the circuit of Fig. 1, let I = total single-phase supply current. Then, including the load resistor previously omitted:

$$I = \frac{E_S}{R} + \frac{E_S}{\dfrac{j\omega L R}{R + j\omega L} + \dfrac{R/j\omega C}{R + 1/j\omega C}}$$

$$= \frac{E_S}{R} + \frac{E_S\{R^2 + L/C + jR(\omega L - 1/\omega C)\}}{2RL/C + jR^2(\omega L - 1/\omega C)} \quad (5)$$

Setting $\omega L = 1/\omega C = R/\sqrt{3}$; $L/C = R^2/3$, then:

$$I = \frac{E_S}{R}\left\{1 + \frac{R^2 + L/C}{2L/C}\right\} = \frac{3E_S}{R}.$$

The total single-phase supply current is thus three times that in each separate load resistor and the magnitudes of the line, reactor and resistor currents are in the ratios $3 : \sqrt{3} : 1$. Under these conditions, the system power factor is unity.

It is to be noted that balanced phase voltages are only obtained at one specific load, corresponding to a particular choice of convertor reactance. A method will be described whereby the desired match between load resistance and phase-convertor reactance can be automatically maintained in the case of varying, symmetrically-balanced loads. Though not of much practical interest, single to two-phase conversion is possible, using a similar circuit but with differently proportioned reactances, so chosen that $\omega L = 1/\omega C = R$.

Saturable-Core Reactors

It is well known that the impedance of iron-cored coils is a function of both the d.c. and the a.c. excitation. A saturable reactor or transductor is a device in which this effect is exploited. The reactor frame carries separate d.c. and a.c. coils, while the construction is such that the two sets of windings are decoupled electromagnetically in such a way as to minimize the mutual inductance.

Fig. 2 shows a typical family of curves which illustrate the performance of a reactor of this type when the standing magnetic bias is altered by variable d.c. excitation.

From curves of this type it is possible to calculate the reactor impedance and, by trial and error, to select an operating point about which equal changes in d.c. excitation cause approximately proportional changes in impedance, remembering that an increase in d.c. bias causes a fall in reactor impedance. A characteristic which is the inverse of this may be secured by winding a second d.c. coil on the reactor core, energizing it from a constant potential source and arranging the connections so that the polarization due to the second coil opposes that due to the first. When the opposing ampere-turns are equal, the impedance is a maximum. Both types of characteristic are required in the convertor network to be described.

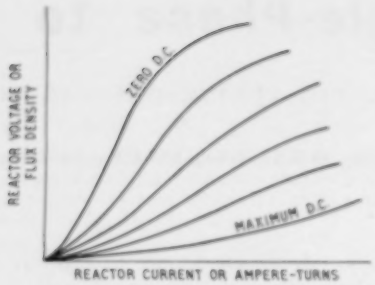

Fig. 2. *Transductor characteristics*

As regards preferred core-shapes, the simplest type which permits the complete isolation of d.c. and a.c. functions is shown in Fig. 3 and is used in one of the units under discussion. A better arrangement is to use twin single-phase cores with the induced a.c. voltages cancelled by proper connection of the d.c. windings as in magnetic-amplifier practice.

The a.c. windings may be joined in series or in parallel.

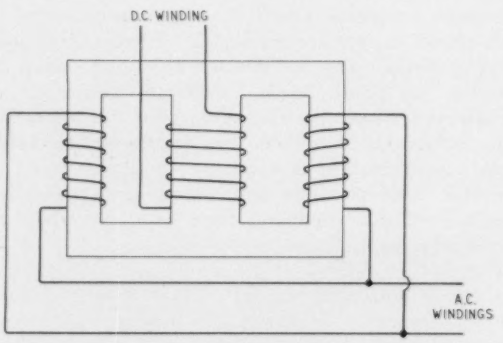

Fig. 3. *Transductor core and windings*

When the connections are properly made there will be no fundamental-frequency induced alternating voltage in the d.c. coil. In general, parallel connection is to be preferred except when the time of response to changing d.c. excitation is required to be extremely rapid. The objection to series connection is that second-harmonic voltages of high amplitude may be induced in the d.c. windings.

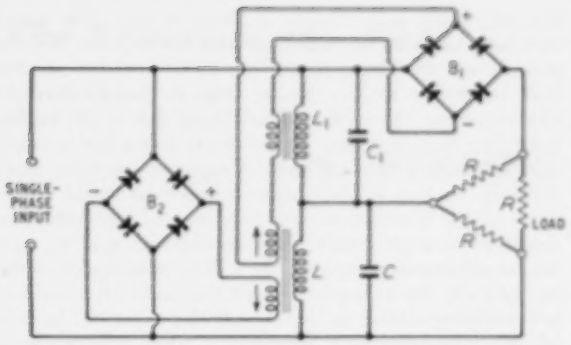

Fig. 4. *Practical convertor circuit*

Referring to Fig. 4, L and L_1 are two saturable reactors, energized by d.c. from the bridge rectifier B_1, connected in series with one line of the polyphase output. Rectified load current will cause partial saturation of L_1 and its impedance will decrease with an increase in the 3-phase line current. The function of the capacitor C_1 will be described later. In the case of the reactor L a second bridge rectifier B_2 (with an associated transformer if desired) is used to provide a preset d.c. magnetization opposing that due to the winding energized from B_1. The net result is that the alternating current in L decreases as the main load current increases. In parallel with L is a fixed capacitor C, drawing a current which is in phase opposition to that in L. The combined current is the difference between the separate branch currents. At no load, neglecting reactor dissipation, the resultant current should be zero. In practice, it is very small. At full load, the magnetizing ampere-turns due to the two coils are equal and opposite so that the reactor impedance is a maximum. By proper design, the full-load reactor current can be made very small and under these conditions the resultant current is practically that drawn by the capacitor alone. If desired, the rectifier B_2 may be connected in series with the capacitor C shown in Fig. 4.

In this manner the desired impedance characteristics of the phase-splitting circuit elements are obtained; i.e., L_1 becomes a variable inductor and the combination of L and C simulates a variable capacitor. The reactances of both the variable elements can be made proportional to the output load resistance. It will be found in practice that the relationship between the exciting ampere-turns and the impedance of a saturable reactor is non-linear at low values of alternating flux density and d.c. excitation and it is desirable to avoid operation over this portion of the characteristic by reducing the number of turns per volt on the a.c. coils. A difficulty is immediately encountered since the maximum inductance of the reactor, with a high value of alternating flux, then becomes insufficient to satisfy the requirements for balanced three-phase voltages at the output terminals of the convertor under light-load conditions. It is possible to reduce or to cancel the high initial lagging current in the reactor L_1 by shunting it by a capacitor C_1 drawing a leading current of such a magnitude that the resultant falls to the value required for phase balance. Normally, for a 1-kVA, 220-V 50-c/s convertor, C_1 will be about 16 μF. A similar correcting capacitance is required in parallel with L, which merely calls for an increase in the size of C.

It is difficult to give exact winding specifications and core dimensions for convertors of this type. For a 220-V unit giving 1 kVA output, using reactors assembled on 3-phase transformer cores of about 2-sq. in. cross-sectional area, roughly 500 turns of wire will be required

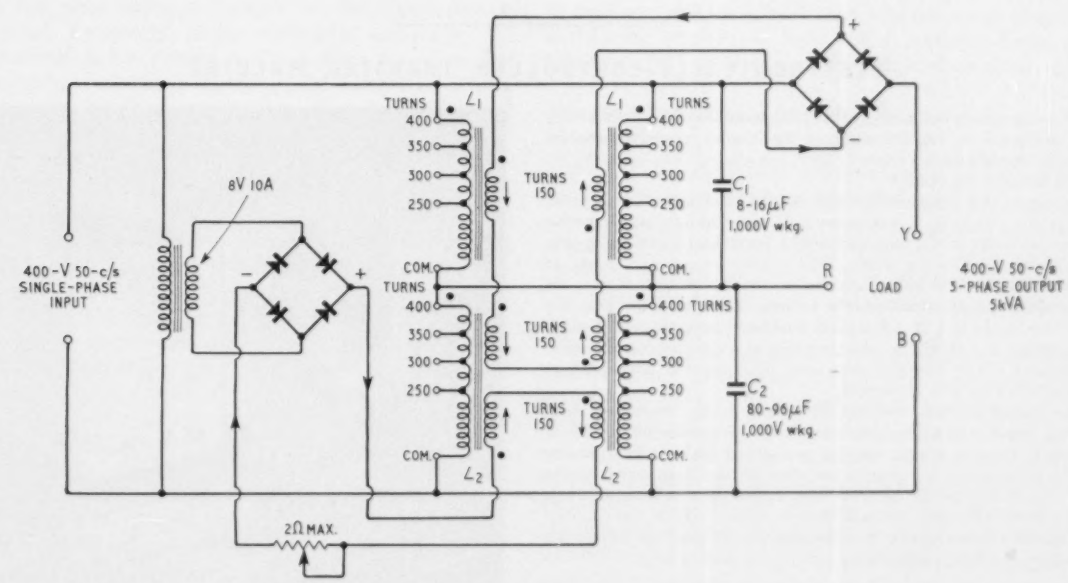

Fig. 5. *Complete circuit diagram of 5-kVA convertor. Reactor specification: All cores, size No.* 41 *laminations* 2½*-inch stack* (4*-off*)*; all windings, No.* 14 *s.w.g. enamelled wire each layer interleaved with* 0.014*-in. paper, finished with Synobel varnish. The numbers of turns and the tapping points are shown in the diagram*

on each a.c. coil, assuming that the two will be parallel-connected. The same number of turns will be required on each series-connected d.c. coil of the two reactors. The ampere-turns on the second d.c. winding on L should be equal to the full-load ampere-turns on the series coil. The capacitance C should be about 80 μF for a 1-kVA, 220-V-50 c/s convertor, the precise value depending on the load power factor and on the initial current drawn by the reactor L. Coarse adjustment of the two derived phase voltages is best arranged by selection of capacitor sizes and by taps on the reactor windings. Auxiliary fine control is then possible by variation of the d.c. excitation of each reactor.

Three-phase cores are not ideal for this type of work and a much better arrangement is to employ two pairs of single-phase cores, each carrying both d.c. and a.c. windings on the centre limb, connections being so arranged that there is no net induced alternating voltage in the d.c. coils.

Performance

A small convertor (output 1-kVA) has been constructed and employed to run a 3-phase motor. The measured phase voltages, under different conditions, are as follows: (a) open circuit: 230 V; (b) motor running light: 220 V; (c) full load: 220 V. In each case, balanced voltages are maintained. The starting torque is good and the line current is relatively low so that direct switching of motors on to the line is permissible. If a number of machines are to be operated from one convertor, it is preferable to start the first motor on light load. Succeeding ones can then be brought up to speed without difficulty, even against a heavy load.

The convertor losses include iron and copper loss in the reactors, rectifier loss and energy dissipation in the d.c. magnetizing windings. For a 1-kVA convertor the d.c. excitation requirement is about 40 W. The metal rectifiers have an efficiency of about 70% (or 90% if germanium or silicon diodes are used) and the reactor core losses are rather greater than in transformers of similar rating, the increased loss being due to the higher operating flux densities. The copper losses are normal.

A second convertor of 5-kVA capacity has been built (see Fig. 5) and, in fact, there appears to be no upper limit to the permissable physical size of equipment of this type, except possibly for troubles caused by harmonic circulating currents. It is sometimes worth while to tune out the prominent third harmonic by means of an auxiliary reactor in series with the capacitor branch of the phase-splitting network. This refinement is included in certain phase convertors developed by Westinghouse Ltd. The output waveform is then extremely good, provided the single-phase supply voltage is sinusoidal.

Reversed Operation

It is sometimes desired to supply heavy single-phase loads (e.g. welding equipment) without unbalancing the three-phase mains. The convertor described is reversible and permits this operation to be performed more cheaply than by using rotary machines. It is possible that the same principles could be applied to the development of a static balancer for polyphase operation.

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ELECTRONICALLY-CONTROLLED TRANSFER MACHINE

A low-cost electronically-controlled transfer machine which has been designed to replace humans for single repetitive transfer assembly operations has recently been introduced into the U.K. by U.S. Industries Inc. Ltd.

Known as the Transferobot 200, it is a machine which is fitted with jaws to pick up a workpiece and transfer it to any position within the limits of the movements of a horizontally-extending arm and a vertically-moving head. The maximum forward stroke of the extending arm is 10 in. and the minimum stroke is 3 in.; the maximum vertical stroke of the moving head is 2 in. and the minimum stroke is $\frac{1}{8}$ in. A typical positional accuracy attained by the machine is $\pm$ 0·002 in. when working at a speed of one operation per second. The machine jaws can fully support and position workpieces of up to 1 lb. weight.

The electronic unit controls the drive to the extending arm, moving head and jaws, and comprises a number of transistor switching circuits which can be pre-set or programmed to any switching sequence by manual switches on the front panel of the unit. To enable two or more of these machines to operate together, each control unit can provide output signals, which indicate the position of the head-piece, and thereby can be used to initiate the operation of other machines.

This picture shows a Transferobot, with a specially adapted head, being used to oil complete clock assemblies. In this application the machine oils eight bearings simultaneously each second

THE DYNAMIC ELECTROMETER

The September article discussed the possibilities as a ballistic galvanometer of an instrument that had not been specifically designed for the job. There is no particular virtue in doing this kind of thing. But it is useful to be able to stretch somewhat limited resources and to know how far and with what sort of penalty they can be extended.

There is, of course, the additional need for this sort of thing nowadays, for teachers who are following the lines of "The Teaching of Electricity" (John Murray), or similar treatments such as those used in many places in the United States and over the whole of New Zealand. Briefly, they begin electromagnetism with the idea of the electric current as a generator of magnetic flux, and this raises at once the problem of measuring Φ and B with at least sufficient accuracy for demonstration purposes. If you cannot do that, indeed, you are no better off than with the orthodox treatment starting with point-poles and inverse-square laws; for, although the poles were non-existent, you could at least locate them and measure them, more or less. We have moved a long way since the days when a speaker at the Institution of Electrical Engineers could solemnly denounce the rationalized m.k.s. system as "a dustbin for unwanted constants"; and this progress would have had a very rakish angle if it had not been possible to justify it by simple experiments.

A Digression on Constants

At this stage in the proceedings, your old man (caparisoned in the livery that you made such a song about) turns up carrying a large dustbin. It is the one they have kept at Savoy Place ever since that memorable meeting, as they can't have people discarding constants all over the premises and cluttering them up. You open it casually expecting to find, perhaps, a few trifles like J, $O^{16} = 16$, and a tenth of a Vernon Harcourt pentane lamp; and a lot of innocent little electromagnetic factors that have been lured there by the Pied Piper (really 4-pied, but it's actionable to call anyone a four-symbol man). To your horror, you discover that the only tenant is the velocity of light, c; and, with this to contain, there just isn't room for anything else. For what the rationalized m.k.s. system does is to prescribe the value of the ampere, so that the permeability of space μ_0 can be *measured*; and furnish you with all the equipment for the *measurement* of the permittivity of space ϵ_0. So that there is no special place in the system for the quantity $(\mu_0\epsilon_0)^{-\frac{1}{2}}$. This is no slight on the status of c, the most wanted of all constants, of whose true-fixed and resting quality there is no fellow in the firmament (unless the satellites of Jupiter really *have* been messing things up, as we were wondering last June). It simply means that the formerly ubiquitous c has disappeared from electromagnetic formulae, in its capacity as a conversion factor. I remember that this disappearing act was cited as an unsoundness in the whole m.k.s. teaching programme by a critic a year or two ago. His objection was that, if you didn't take $\mu_0 = 1$ and $\epsilon_0 = 1$ (which, of course, you never do simultaneously!) you had to obtain a value for ϵ_0 by using c^2 and μ_0, with no justification of this process to begin with, and with a thankless jettisoning of c^2 thereafter. If this was what people actually did in their teaching, he would have been quite right. But they don't. They measure ϵ_0 from first principles. And this is the second really decisive experimental step.

This all sounds happily dogmatic. It isn't meant to be. I have been trying to read O'Rahilly recently, and he would say it's all wrong anyway—in the sense that μ_0 and ϵ_0 don't exist, and that whatever the factor c is it isn't the velocity of light. But, without being dogmatic, we have to measure the value of what we *call* ϵ_0.

Simple Capacitance Measurements with a Microammeter

We will skip the theory which, starting with Coulomb's law $F = Q_1Q_2/4\pi\epsilon_r\epsilon_0 r^2$ leads to $C = \epsilon_0 A/d$ for a parallel-plate air capacitor and $C = 4\pi\epsilon_0 R$ for an isolated sphere of radius R metres in air. The only thing we are worried about is measuring these values of C in a way that beginners can understand, and obtaining a value of ϵ_0 that they will accept.

The circuit is the usual charge-discharge system, incorporating a vibrator which runs from the 50-c/s mains. An electrically-maintained fork would probably be better for the job, and I have used one of these; the question here is really whether it is worth while to accept a higher capacitance in order to have the higher frequency. The simple vibrator unit that Philip Harris supplies has a capacitance of 15–20 pF, effectively in parallel with C, which is of the same order as the capacitances you want to measure, and larger than many of them. Forget about the capacitance of the vibrator unit, though, for the moment. It can easily be eliminated in cases where it is significant. Let us suppose that you make up an air capacitor, with square 20-cm sided

plates separated by, say, 3 mm; then $C = \epsilon_0 A/d = 8.85 \times 10^{-12} \times 4 \times 10^{-2}/3 \times 10^{-3} = 1{\cdot}2 \times 10^{-10}$ farad. If $V = 100$ volts and the vibrator frequency is 50 c/s, the mean current in coulombs per second is 6×10^{-7} amperes. For a meter with a sensitivity of $4{\cdot}5 \times 10^{-8}$ amperes per millimetre, the reading will be nearly 13·5 mm. I remember, as it happens, doing the experiment in this way using a 128 c/s fork to work the vibrator, being able to show that the capacitance is proportional to the overlap area A and to $1/d$, obtaining a reasonable value for ϵ_0, and also finding ϵ_r for Perspex. It isn't, of course, highly accurate, and there is the snag about the capacitance of the vibrator that didn't occur to me at that time as important. I suppose that most people do this, and there would be nothing worth reporting if I hadn't had my attention drawn to a commercial model from the same firm, which is fifteen times as sensitive—$2{\cdot}9 \times 10^{-9}$ ampere per millimetre, to be precise. This puts the whole experiment right down into the picofarad region.

Unless you belong to the old school and were brought up to make measurements in jars, you probably remember instinctively that the capacitance in picofarads of an isolated sphere is roughly the same as its radius in centimetres. If, then, we have a milli-microammeter meter that can play with picofarads, we can do the kind of experiment that is usually done as electrostatics, but do it *dynamically*, and quantitively. The milli-microammeter is a possible *electrometer*.

The Differentiating, or Dynamic, Electrometer

The ballistic galvanometer is really an integrating ammeter. The microammeter used with a vibrator is a differentiating electrometer. One measures Q as $\int I dt$; the other measures I as dQ/dt. The high sensitivity of the meter I have mentioned is, for some purposes at least, bought at a high price, for the resistance is 1,400 Ω and the critical damping resistance 25,000 Ω. But, for differentiating purposes, the resistance does not matter at all. Our only concern is with the current sensitivity. The capacitance of the meter itself is also immaterial.

Isolated Spheres

This record of the kind of experiment that can be done may be interesting, particularly as it revealed some snags which I haven't yet resolved. Spheres A (R = 5.3 cm = 0.053 m) and B (R = 6·5 cm = 0·065 m) were mounted "isolated spheres" straight out of the electrostatics cupboard; C (R = 11·7 cm = 0·117 m) was the best the chemists could do in the way of flasks, carefully covered with aluminium foil—but it wasn't completely spherical.

Contact was made to the surface of the spheres at the point corresponding to X in the circuit of Fig. 1. Since they are supposed to be "isolated"—that is, the earthed side of the capacitor of which they form the "charged" side is the rest of the lab.—no other connection was made, of course. In this situation, the sphere-earth capacitor is in parallel with the vibrator-earth capacitor.

The graphs of meter current against p.d., for a 50-cycle vibrator, are shown in Fig. 2. I was a bit lazy about B, but followed the rest up at 30, 60, 90, 100 and 110 volts.

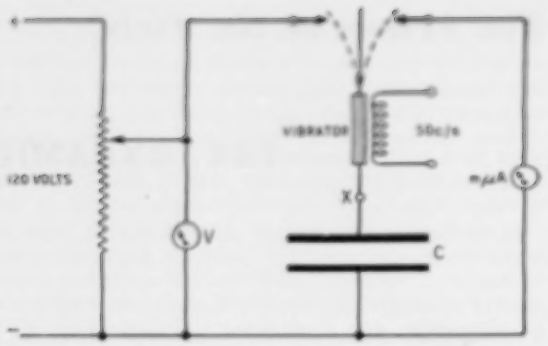


Fig. 1. The stock vibrator circuit for dynamic capacitance determinations. The microammeter is really a milli-microammeter, the Pye Scalamp 7904/T. The vibrator runs from the 50-c/s mains. X marks the spot to which the surface of the isolated sphere is connected; when that happens, the vibrator acts as a capacitance in parallel with it

Looking at the three that have been followed all the way, the first thing that has been demonstrated is that the systems all really do have a capacitance. As $I = nQ$, where n is the vibrator frequency, if I/V is constant, then Q/V is constant. This is not just a bit of theory that follows from the inverse-square law and the concepts of electric flux-density and field-strength and potential gradient. It is as concrete a piece of experimental work as Ohm's law is. And, of course, if we accept the calibration of both meters at their face values, it is just as basic; no other law or result has been assumed in order to obtain it.

The dotted line, at $V = 100$ volts, has been put in to simplify the arithmetic. As $Q = I/n$, $Q/V = I/nV$ and, for the 50-cycle vibrator, nV is then 5,000. I have to keep things at this sort of level, otherwise the letters pour in from all the computing organizations pointing out that I've mislaid a decimal point or lost Boltzmann's constant or just guessed instead of estimating, or that I ought to devote my talents to politics. Divide the value of I on the graph for $V = 100$ volts by 5,000, then, and you get C in farads. We thus have:

Vibrator alone	$13{\cdot}6 \times 10^{-12}$	farad
Vibrator + A	$18{\cdot}0 \times 10^{-12}$	,,
Vibrator + B	$20{\cdot}0 \times 10^{-12}$	,,
Vibrator + C	$31{\cdot}6 \times 10^{-12}$	,,

Giving for the spheres A, B and C:

A,	measured capacitance,	$4{\cdot}4 \times 10^{-12}$	farad
	calculated value from R,	$5{\cdot}9 \times 10^{-12}$	,,
B,	measured capacitance,	$6{\cdot}4 \times 10^{-12}$	,,
	calculated value from R,	$7{\cdot}3 \times 10^{-12}$	,,
C,	measured capacitance,	18×10^{-12}	,,
	calculated value from R,	13×10^{-12}	,,

You may smile a little at these results. I did—not because of their obvious crudity, but in the sheer satisfaction of having got the thing to work at all in this region. For A and B, which are much smaller than the capacitance of the vibrator, there seems to be nothing that a little more care and trouble in the routine could not straighten out. But the big chap C, which errs in the wrong sense, raised a new problem that I have not yet solved completely.

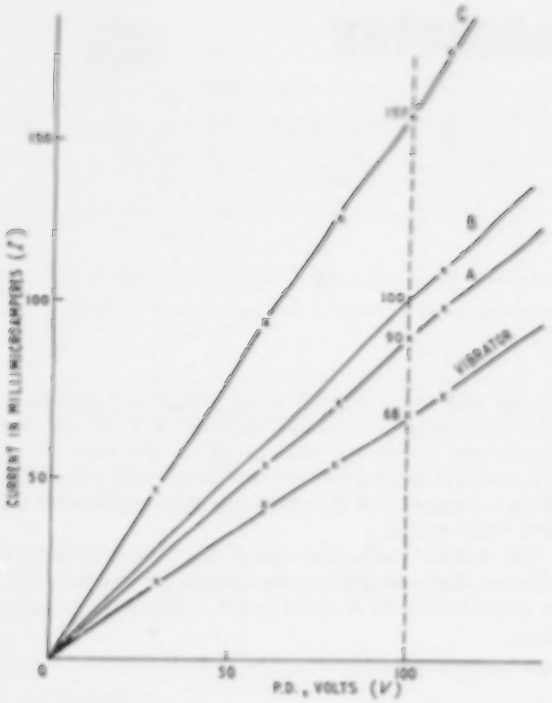


Fig. 2. Current-p.d. graphs for vibrator alone, and for vibrator with spheres of various radii. That is, A for vibrator plus A, and so on

How Isolated and how Spherical must an Isolated Sphere be?

That this is indeed a genuine capacitance experiment, and not the product of leakage over insulation (which at 100 volts oughtn't to be difficult anyway) is easily shown by bringing your hand up to the spheres. In fact, with C, the spot could be deflected right off the scale and, with A, a simple demonstration of the concentric-sphere capacitor resulted if one brought both hands close round it with the usual butter-fingers gesture of a cricketer.

It seemed likely, therefore, that while the odd 20 cm of polythene rod was all right for A, C was simply not isolated enough; so it was mounted well away from the bench on a metre of glass tubing. This brought the capacitance down to 16×10^{-12} farad, which differs from the calculated value by about the same as the others, but is still on the wrong side. I don't think it could have been isolated much more, at least in this room. What I have been wondering is whether the small departure from the sphere due to the flattened standing-on part could have made all that difference. A small toupee of foil, rather unskilfully attached, raised the capacitance even more! Is it a job for the mathematician, or ought I to do things with much greater care? One thing that does seem clear is that, if one is going to use "electrostatic" apparatus *on the bench* for this kind of thing, it isn't much use scaling up the dimensions. I know it's a misquotation, but the earth is too much with us.

Conclusion

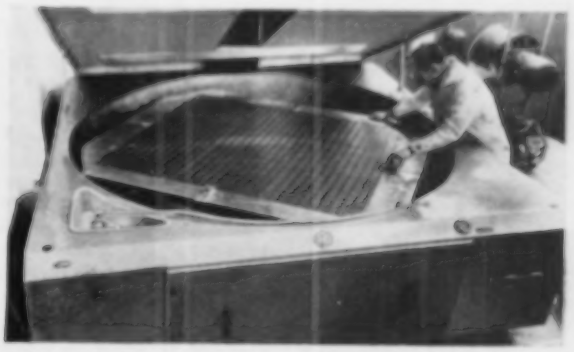
What use has this exploration of the milli-microammeter been? Well, I think it has brought a quantitative element into a couple of basic experiments, those with the parallel-plate capacitor and the "isolated" sphere, that does not usually accompany demonstrations with the *static* electrometer. It should be easy enough to devise simple drills on capacitors in series and in parallel, too. I think that the next step is to see how a faster vibrator behaves; but I shall have to do a good deal more thinking before I commit myself to any more writing.

AUTOMATIC PRINTED-CIRCUIT ETCHING

Printed Circuits Ltd., of Boreham Wood, one of the manufacturing concerns specializing in the production of printed circuits, name-plates and instrument panels, are using two completely automatic machines for etching and anodizing.

These each comprise a number of tanks joined together to form a large circular assembly, agitators, a centre circular structure with projecting arms which support the work-piece to be etched or anodized, and a console which houses electronic equipment for automatic control of the variable factors in the process. In operation, the various tanks are filled with the appropriate agents such as etchants, degreasers, neutralizers and the process programme is set into the control console to control the time-cycling and also the chemical contents and temperatures of the processing tanks. As soon as the work-pieces are clamped to the carrying arms and the machine switched on, the required cycle of events start and continue automatically until the finish. When processing large quantities, the machine is set to cycle continuously and an operator loads and unloads the work-pieces into the carrying arms.

This shows one of the automatic etching and anodizing machines with the control console in the background



Illustrated here is another automatic processing unit for 'developing' master plates of up to 25 sq. ft.

Binary-To-Decimal Translater

USING A MODIFIED ELECTRONIC COUNTER

By D. A. German, B.Sc.(Hons.)

There is a growing need for a small and simple binary-to-decimal decoder as many measuring techniques are considerably simplified in the binary form. One well-known example is that of weighing, using a chemical balance. The weights required are one only of each size in the series 2^n; i.e., $1(2^0)$, $2(2^1)$, $4(2^2)$, $8(2^3)$, $16(2^4)$ . . . . $32768(2^{15})$, $65536(2^{16})$, . . . etc., units. For weighing to, say, five decimal significant figures, this requires 17 weights. The total weight is now in binary form and requires translation to decimal form as it is completely unintelligible to the ordinary mortal. With a quick and automatic translater, the process would be simpler than by present methods. In electrical engineering, binary potentiometers have been in existence for a long time and they, as well as digital voltmeters working in the binary scale, require a translater. In tracing faults in a digital computer, there may also be application for a separate unit to translate binary numbers up to 33 digits.

Most methods of conversion are only suitable as computer routines, being mathematical processes to convert the binary figure to the binary coded equivalent. Other methods require the use of a decimal desk calculator and require either a knowledge of decimal equivalents or a liking for repeated simple calculations. The method in general use is to add the decimal equivalents of the binary digits; e.g., for

$$1\,0\,1\,0\,1 \equiv 1\times16 + 0\times8 + 1\times4 + 0\times2 + 1\times1 = 21$$

This seems quite simple for calculations of this size but, for numbers greater than 10 binary figures (3 decimal figures), it becomes far too exhausting.

The decoder envisaged by the author requires that an existing decade counter employing Eccles-Jordan circuits be made to add the decimal equivalents by means of gating circuits.

Principle

The counter to be employed is that of the 'eight + six' type, a block diagram of which is given in Fig. 1. This is a well-known method of making a counter recycle on the count of 10 using a circuit which consists of four Eccles-Jordan triggers with a feedback loop. Normally, without this loop, the circuit would recycle on the 16th input pulse, as each stage gives an output on its second input pulse. To make the circuit recycle on the count of 10, a method must be found to add 6 into the count at some point in the cycle. This can be done by using one or more feedback loops. The 'eight + six' counter uses two feedback loops which inject pulses into the second and third triggers when the count reaches 8. This is equivalent to adding a 2 and a 4 (6) into the circuit and brings the counter up to the state of 14. The count then progresses normally up to the state of 24 when it recycles and thus gives an output pulse after the 10th input pulse.

Put another way, the count may be described as follows: Each trigger can be in two states which may be described as the 0 and 1 states. The count proceeds thus:

Count	State of counter		Decimal state of counter
0	0 0 0 0		0
1	0 0 0 1		1
2	0 0 1 0		2
3	0 0 1 1		3
4	0 1 0 0		4
5	0 1 0 1		5
6	0 1 1 0		6
7	0 1 1 1		7
8a	1 0 0 0	feedback + 0 1 1 0	8
8b	1 1 1 0		14
9	1 1 1 1		15
10	1 0 0 0 0	output and recycle	0

There are several other methods of recycling the circuit at the count of 10 but this one appears to be the only circuit suitable for the proposed decoder. If fuller details are required on the operation of this and other counter circuits, they may be found in most books on pulse circuits.

The state of the count must be indicated and the usual method is to use neon indicator lamps connected across the anode loads. This gives the state of the count in binary form and, in order to be able to read the state in decimal form, a more complex circuit in the form of a matrix is necessary. Fig. 2(a) gives one suitable read-out matrix. For the purposes of the following description

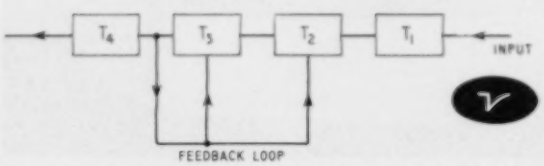


Fig. 1. *Block diagram of an 'eight + six' counter employing four Eccles-Jordan trigger circuits*

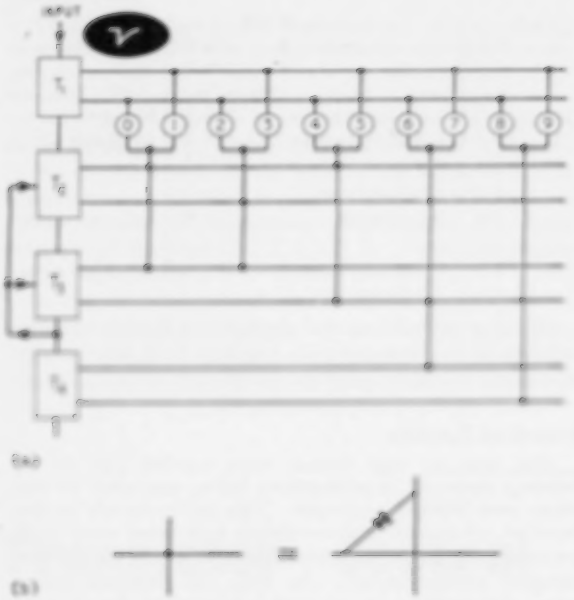


Fig. 2. *The decoding matrix for an 'eight v. six' counter employing diodes and neon indicator lamps*

of its operation, the junctions will be considered to be of the type shown in Fig. 2(b).

It will be noticed that any vertical line up to a pair of indicators constitutes an 'AND' gate. This means that unless both the inputs to the gate are at a low potential, the output will remain at a high positive potential. The other side of each of the indicators is connected to one of the anodes on the first trigger as shown. When one of the gate outputs is at a low potential, there will be a potential across one of the indicators as one of the anodes of the first trigger will be at a high potential. There will also be a potential across one of the indicators when the output is at a high potential but, by taking the connections to the first trigger to some point along the load resistors instead of to the anodes, it is possible to reduce this potential to some value below the firing potential of the neons. In this way, the first trigger is used to determine whether the number is odd or even.

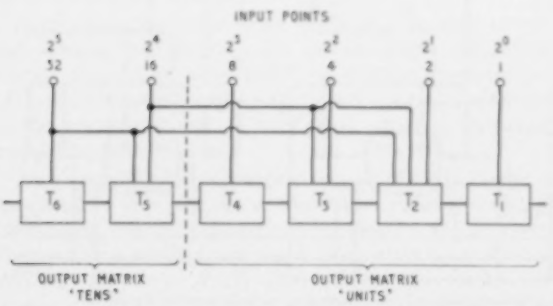


Fig. 3. *Block diagram for the trigger-circuit inputs for a decoder of five binary figures*

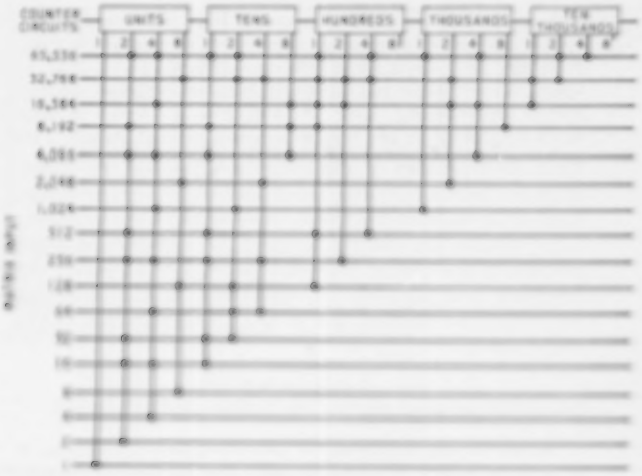


Fig. 4. *Input matrix for a five-decimal figure decoder employing 'eight v. six' counter circuits*

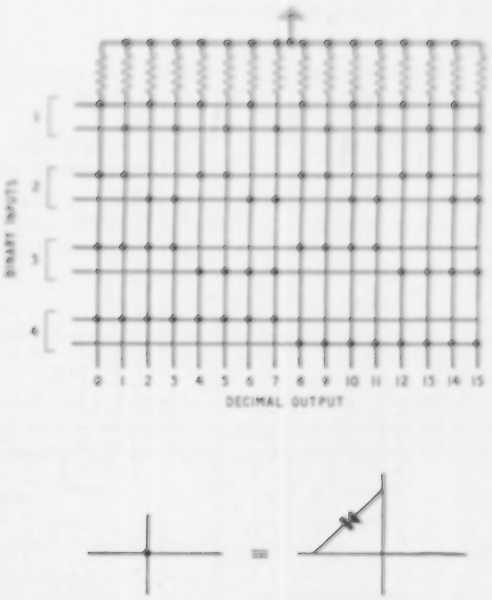


Fig. 5. *A complete decoding matrix to decode four binary figures*

The other stages determine the remaining factors of the count by means of the matrix of 'AND' gates.

In practice, resistors may be used instead of diodes in the gates, though it makes the gates less positive in action.

As can be seen from the description of the counter circuit, a 6 is injected into the circuit in parallel form and the author could see no reason why it should not be possible to do this with other numbers. It was particularly hoped that it would be possible to inject the decimal equivalents of the binary digits in this way and so obtain the sum of these for any binary number.

A matrix of the form shown in Fig. 3 is required to apply the input pulses to the correct trigger circuits. A larger and more glorified version is given in Fig. 4. Each binary digit is converted into its decimal binary coded equivalent; e.g., $2^5 = 3, 2 = 1\,1\,0\,0\,1\,0$, and is then applied to the trigger circuits in that form. The main disadvantage of this method is that all the inputs must be read in turn and not simultaneously. The inputs must all be isolated from each other by diodes or switches.

To enable the counter to carry out the ensuing summation, it was found necessary to modify it so that the feedback was applied to both grids of the receiving Eccles-Jordan circuits, and to put a delay in each of the carry circuits.

Theoretical Results

As all decade stages carry out the same operations, it is only necessary to consider one. For correct operation, it must be possible to add any number from 1–9 to any number that the counter can hold. This is carried out successfully by the above circuit, and a few examples demonstrating the varying degrees of difficulty and the necessity for delays are given below:

2+4		3+2	
2	0 0 1 0 in counter	3	0 0 1 1 in counter
4	0 1 0 0 add	2	0 0 1 0 add
6	0 1 1 0 sum		0 0 0 1 logic sum
			0 1 0 0 delayed carry
		5	0 1 0 1 sum

5+4		4+8	
5	0 1 0 1 in counter	4	0 1 0 0 in counter
4	0 1 0 0 add	8	1 0 0 0 add
	0 0 0 1 logic sum		1 1 0 0 binary sum
	1 0 0 0 delayed carry		0 1 1 0 feedback
	1 0 0 1 binary sum		1 0 1 0
	0 1 1 0 feedback		1 0 0 0 delayed carry
9	1 1 1 1 counter sum 12 binary 14		1 0 0 1 0 decoded sum

In order not to bore the reader with further demonstrations of the decoder's workability, let it just be said that it will in theory carry out the remaining summations.

Practical Results

The tests on this system were carried out on an existing counter, modifications being required in the carry and feedback circuits. The carry circuit in the counter, of course, had no delays and these were built on separate chassis. The circuit used was a set of cathode-coupled multivibrators. At first, the feedback pulses were not sufficiently powerful to trigger the Eccles-Jordan circuits reliably when they were applied to both grids of two of these circuits. A cathode follower overcame this.

The circuit, as described above, functioned correctly, though a certain amount of difficulty was experienced in directing the input pulses via the input matrix equally, and with sufficient power to trigger the Eccles-Jordan circuits simultaneously.

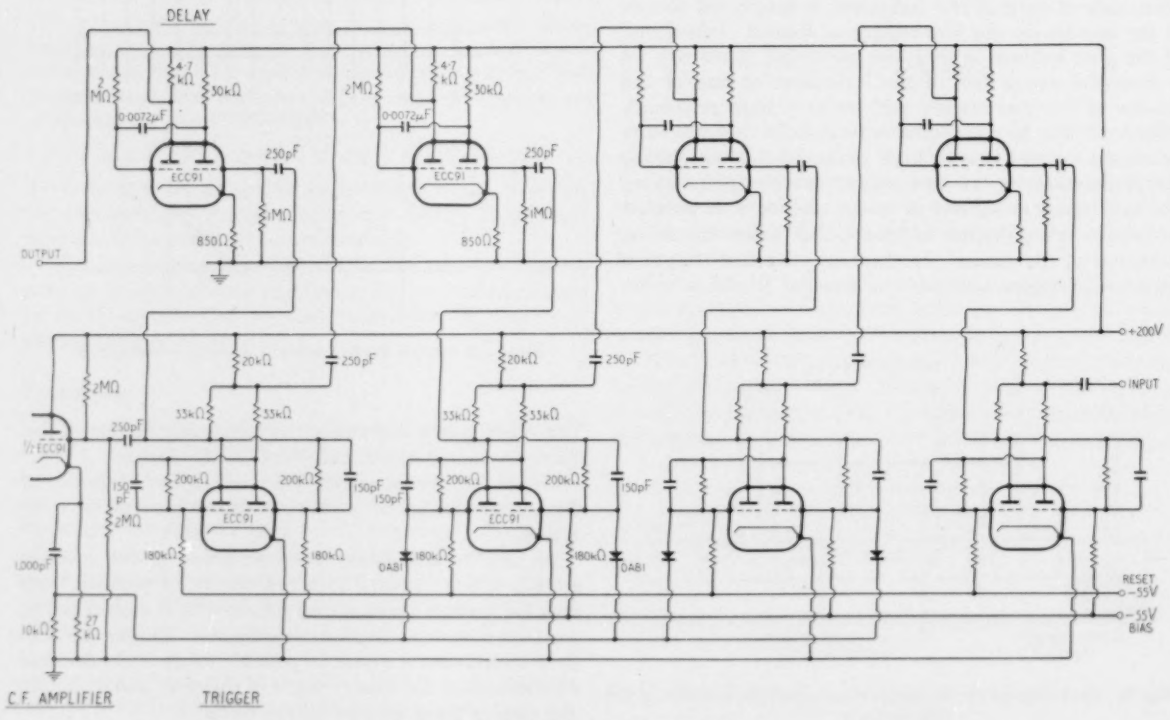


Fig. 6. *The circuit diagram of the modified counter*

Conclusions

Though successful, the modification of an existing counter added to the complexity of the final circuit and it would have been easier to have started from scratch. The merit of this method is that there is no excessive increase in the number of components required as the range of the machine is extended. The only part of the circuit where the increase is not linear is in the input matrix.

The number of components required for the input matrices of decoders of 1 2 3 4 5 & 10 decimal places are given in Table 1.

As can be seen, even for this there is no comparison with the number required for a complete decoding matrix of the type shown in Fig. 5. A decoding matrix for 17 binary figures, 5 decimal figures, requires $17 \times 99{,}999$ or 17×10^5 junctions and 99,999 indicators.

TABLE 1

Number of decimal places	Number of binary places	Number of junctions	Greatest number of junctions to any input
1	4	4	1
2	7	13	3
3	10	25	5
4	14	41	5
5	17	67	10
10	33	286	17

The above figures would appear to be reasonably economical on components.

REFERENCE

I. E. Grosdoff, "Electronic Counters", *R.C.A. Review*, September 1946, p. 438.

Transistor Control of A.C. Generators

By E. L. Robinson*, A.M.I.E.E.

Road travel between two places in London by public transport today takes substantially the same time as it did 100 years ago. The increase in vehicle speed has been almost completely offset by the time spent idling in traffic delays. Today half the operating time of a bus is spent stationary with the engine idling.

To relieve the batteries of this heavy standby demand, the engine-driven generator must carry the load at 'tick over'. This requires generation over a speed range at least as great as 10 to 1. Such a range cannot be achieved by simple field weakening on a d.c. generator without commutation difficulties.

The a.c. generator avoids these difficulties by using a semiconductor rectifier stack instead of a commutator. With improvements in the control unit, satisfactory operation can be achieved with an a.c. generator operating with extremely weak fields, fed with d.c. derived from the rectified a.c. output.

Field distortion with weak fields can be very severe and coarse regulation, as occurs with vibrating contact regulators, may result in voltage transients of sufficient magnitude to damage, if not destroy, the diodes in the rectifier stack. This risk is avoided by using proportional regulation.

With proportional regulation the controlled voltage is permitted to rise above the control or reference value. The excess voltage, sometimes referred to as the error function, is amplified and used degenerately to reduce both the alternator excitation and the rise in output voltage. If V_f is the voltage applied to the field of the alternator to attain an output voltage of V_o at the d.c. terminals of the rectifier, the alternator will have a voltage amplification of n where

$$n = \frac{V_o}{V_f}$$

$$\text{or } V_f = \frac{1}{n} V_o$$

The excitation voltage V_f will be obtained by backing-off the supply voltage V_o with an amplified error function αV_e where α is the regulator amplification. Therefore

$$V_f = V_o - \alpha V_e$$

$$\text{or } V_e = \frac{1}{\alpha}(V_o - V_f)$$

Substituting for V_f we get:

$$V_e = \frac{1}{\alpha}\left(\frac{n-1}{n}\right) V_o$$

$$\text{or } \frac{V_e}{V_o} = \frac{n-1}{\alpha n}$$

Since the control or reference voltage

$$V_c = V_o - V_e$$

$$\frac{V_e}{V_c} = \frac{V_e}{V_o - V_e} = \frac{1}{\frac{V_o}{V_e} - 1}$$

So that

$$\frac{V_e}{V_c} = \frac{1}{\frac{\alpha n}{n-1} - 1} = \frac{n-1}{\alpha n - n + 1}$$

* *Simms Motor Units Ltd.*

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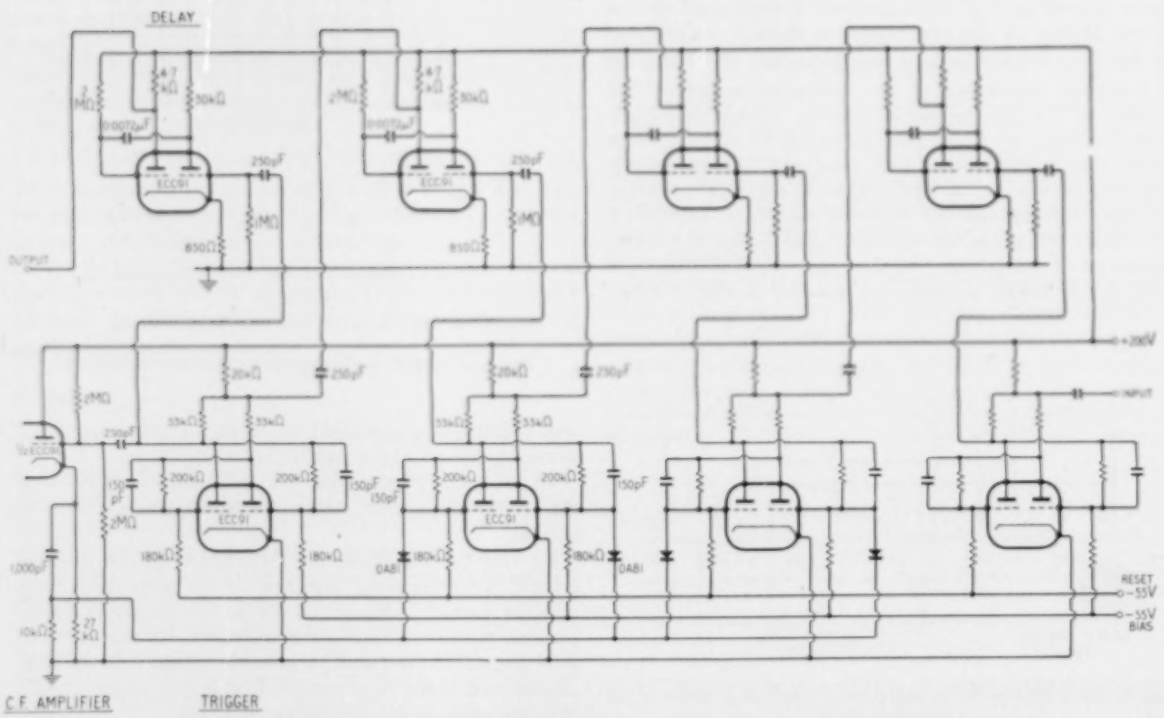

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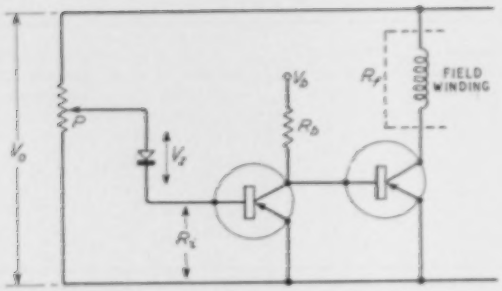


Fig. 1. *The basic circuit diagram*

If α is very large $-n+1$ will be negligible in comparison to αn and $\alpha n - n + 1 \triangleq \alpha n$ therefore

$$\frac{V_e}{V_c} \triangleq \frac{1}{\alpha} - \frac{1}{\alpha n}$$

At high speeds, when n becomes large enough, even the second term may be neglected and $V_e/V_c \triangleq 1/\alpha$.

The advent of the transistor enables amplification to be achieved at low voltages and comparatively heavy currents. This renders it especially suitable for control of vehicle generators. With the transistor came the Zener diode. This device gives an ideal method of separating the error function from the output voltage. Fig. 1 gives the basic circuit diagram of a transistor voltage regulator. This is very simple as may be seen. A potentiometer and Zener diode are used to sample the output voltage and feed a signal to a two-stage transistor amplifier, the output of which supplies the excitation to the generator. The potentiometer is used solely to give a means of setting the output voltage by adjustment of the ratio of the reference level PV_z and the Zener-diode voltage V_z. The signal applied to the input of the amplifier will be

$$\frac{V_o}{P} - V_z = \frac{V_e}{P}$$

where P is the ratio of the potentiometer and V_z the inverse breakover voltage of the Zener diode. The amplifier is designed so that with zero signal input the output transistor is just bottomed. The output voltage developed across the field winding will be

$$V_f = V_o - \frac{\beta R_f}{P R_l} V_e$$

where β is the overall current gain of the amplifier and R_f and R_l the output load impedance and input impedances respectively.

Thus it may be seen that the excitation is equal to the output voltage minus an amplified error function or, as previously stated,

$$V_f = V_o - \alpha V_e$$

Where the amplification term $\alpha = \beta R_f / P R_l$

Using transistors with current gains of the order of 100 with well-chosen operating points and a Zener diode having a breakover voltage close to the desired reference value, it is quite possible to achieve a value for α of the order of 1,000.

As previously shown, the error function V_e even at high speeds is substantially $1/\alpha$ of the reference value which, in theory, could give a maximum error of 0·1% of the reference voltage.

The breakover characteristic of the Zener diode, however, is not absolutely rectangular and an additional error is introduced from variation of the practical reference value PV_z. Variation in the current gains of the transistors has little effect as long as the product of the respective α's is large. The shape of the voltage characteristic with variation of speed will be substantially that of the Zener voltage versus Zener current, characteristic with the current axis calibrated in terms of speed. In the circuit diagram an auxiliary supply potential has been shown in addition to V_o. The import of this has not been explained. The description of the circuit has assumed the existence of the nominal supply potential and of field current to enable the alternator to generate.

When an alternator is arranged to supply its own excitation and run up to speed, there will be, in the first instance, no field current and no resultant generated e.m.f. There will be, however, a small voltage, 200–1,000 millivolts, produced from the remanant magnetism of the iron parts of the generator-field system. For this voltage to become cumulative the local circuit including the machine and the power transistors must have a loop gain greater than unity. The input resistance to the

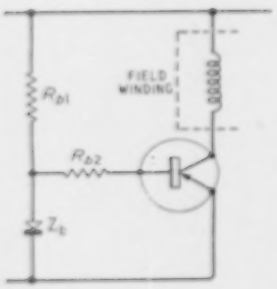


Fig. 2 (*Above*). *The build-up circuit*

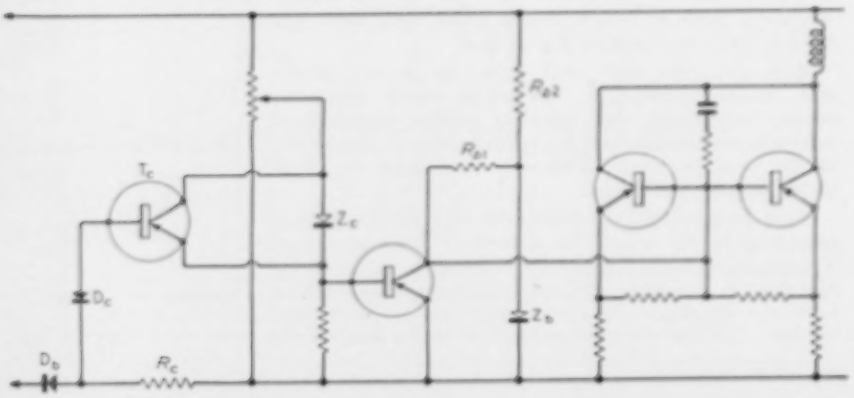


Fig. 3 (*Right*). *Typical regulator circuit*

power transistor is too high for this to be the case, since the circuit parameters are designed for a supply potential around 50 times larger than this residual e.m.f.

This situation is overcome by the introduction of an auxiliary bias regulator employing a large Zener diode. Fig. 2 shows this part of the circuit. The bias resistor has been reduced in value and split into two parts. The breakover voltage of the Zener diode Z_b is about a third of the nominal supply voltage. The direct bias resistor R_{b2} is, therefore, about a third of the resistance of the single resistor of the previous diagram, and will thus give correct bias at normal working. The resistor R_{b1} is low enough to permit cumulative build-up from the low residual voltage. The rating of the Zener diode is chosen to work safely from the nominal supply voltage in series with resistor R_{b1}.

The response time of the regulator is so fast that it will attempt to modulate the field in response to the ripple of the full-wave rectified 3-phase a.c. supply. This is disadvantageous, as it results in high voltages being developed at the collector of the power transistors when they are carrying quite high currents. This represents a considerable increase in power dissipated in the transistors.

This has been overcome by feeding back a signal proportional to the rate of change of field current to the base of the power transistors. It is achieved by linking the collector and base of the power transistors by a resistor and capacitor in series since the voltage transient developed across the field is proportional to the rate of change of current through the field. The capacitor isolates the steady field voltage. The capacitor and resistor must be proportioned to avoid low-frequency oscillation of the system at the resonant frequency of the field winding inductance and the d.c. blocking capacitance. Thermal stabilization is achieved by using transistors capable of considerably more power dissipation than is asked of them. It must be accepted that there will always be a temperature at which thermal run-away will occur but cooling is such that self-heating is negligible in comparison with heating from high ambient temperatures. For use on vehicles which may be used on any duty anywhere in the world there always appears some duty somewhere that exceeds preconceived notions of maximum expected temperature.

To counter this, the risk of run-away of the driver transistor is made greater than that for the power transistor. When the unexpected happens the power transistors will be starved and the system will shut down without permanent damage to the equipment.

When operating power transistors on an inductive load at moderately high supply voltages there is always a risk of transiently crossing the danger line into the avalanche region. In this area the collector current may become self-sustaining. It is characterized by the existence of reverse base current. If a sufficiently low resistance d.c. path between base and emitter is not existent, control of the transistor will be lost. To avoid this happening when the driver transistor is cut off in response to a transient, this transistor is shunted by a resistance limiting the maximum impedance that can be developed.

The shunting resistance will always reduce the source impedance to the power transistor. This will improve the thermal stability of the power transistor circuit.

The avalanche voltage for a transistor falls with increase of collector current. To exploit this, two transistors are used in parallel for the higher system voltages. To improve the load sharing of the paralleled transistors, negative feedback is obtained by inserting small resistances in series with the emitters. This also improves the thermal stability of the transistor power stage.

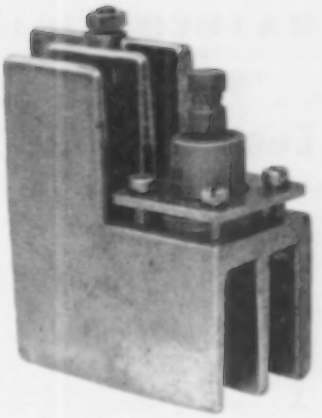


Fig. 4. One of the eight diodes in a rectifier unit showing how each of these is mounted on a heat sink

It was mentioned earlier that the consistency of the regulated voltage is affected by variation of the reference voltage dependent on the control Zener-diode characteristic. This is particularly noticeable at the onset of Zener conduction of this reference diode. This can be improved by increase in the operating current of this diode by shunting the input of the following transistor. This improvement must be a compromise since the lower source resistance will improve the thermal stability of the drive stage which preferably must not be better than the power stage.

When used with alternators the full current is limited inherently by the reactance of the machine. It may at times be desirable to limit the maximum current to some lower value. This may easily be achieved by a simple additional circuit.

To sense the current a small resistance R_c in Fig. 3 is inserted in the main positive d.c. line common to the regulator. The voltage drop at the limiting current across this resistor will be of the order of half a volt. As a reference value to delay the operation of this circuit, use is made of the very non-linear characteristic of the voltage drop in the forward direction of a small germanium diode. The voltage signal obtained from the low resistance R_c is fed through this diode D_c to an additional transistor T_c short-circuiting the control Zener diode Z_c when this transistor conducts. Thus a rapid reduction of system voltage occurs with increase of current above the predetermined level. In some cases the reference diode may be omitted, the delay is then obtained from the threshhold voltage necessary before the emitter diode of the current-control transistor T_c will conduct.

A final circuit with these refinements is shown in Fig. 3. Transistor control is not limited to use with a.c. generators and is just as useful with d.c. generators. In such cases it may be combined with the semi-conductor equivalent of the orthodox cut-out which need only be a large blocking diode D_b in Fig. 3. A blocking diode may be used between rectifier and battery on an a.c. generator system when the auxiliary uses of a cut-out must be maintained.

Feedback Analysis using the Root-Locus Method—2

In our last article, we showed how to determine graphically points where the expression

$$1 + A\beta = 1 + K\frac{N(p)}{D(p)} = 0 \quad .. \quad .. \quad .. \quad (1)$$

when $N(p)$ and $D(p)$ are polynomials in p, and K is a constant. The remarkable fact about the root-locus method is that it is easier to determine the complete curve whose points are the roots of Equ. (1) when K is allowed to vary than to determine the particular points on that locus associated with a given value of K. Particular attention was devoted to an example in which $N(p)$ was linear in p and $D(p)$ was cubic, but as this example did not have all the features which commonly occur, we shall now consider that of a three-stage RC amplifier in which

$$A\beta = A_0\beta_0/[(1 + p\,T_1)\,(1 + p\,T_2)\,(1 + p\,T_3)] \quad (2)$$

after which we shall be in a position to understand certain general rules which simplify the determination of root-loci. We shall suppose that the numerical values of T_1, T_2 and T_3 are respectively 2, 5 and 10 microseconds, and that the unit for p is 10^5 radians/sec. If we then let $K = A_0\beta_0/T_1\,T_2\,T_3$, the equation whose roots determine the closed-loop behaviour of the three-stage RC amplifier becomes

$$1 + A\beta = 1 + \frac{K}{(p+1)\,(p+2)\,(p+5)} = 0 \ldots \quad (3)$$

so that, in the notation of Equ. (1),

$$N(p) = 1, \quad D(p) = (p+1)\,(p+2)\,(p+5) \ldots \quad (4)$$

There are therefore no zeros, and three poles at -1, -2 and -5 respectively; if these are plotted in Fig. 1 and X or $\alpha + j\omega$ is any point of the root-locus, and X is joined to the three poles so that $XP_1 = R_1$ and XP_1 makes an angle θ_1 with the positive direction Ox in the sense indicated by the arrow and R_2, θ_2, R_3, θ_3 are defined similarly, we have, as the condition for X to be on the root-locus

$$\theta_1 + \theta_2 + \theta_3 = 180^\circ \quad .. \quad .. \quad .. \quad .. \quad (5)$$

$$R_1\,R_2\,R_3 = K \quad .. \quad .. \quad .. \quad .. \quad .. \quad (6)$$

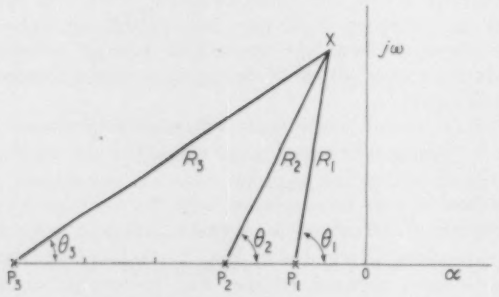


Fig. 1. *A general point of the root-locus associated with Equ.* (3) *and its relation to the poles*

If, as before, we first seek places where X can be on the real or α-axis, it is clear that X can either be to the left of P_3 [θ_1, θ_2 and θ_3 all 180°], or between P_1 and P_2 [$\theta_1 = 180^\circ$ and $\theta_2 = \theta_3 = 0^\circ$]. But this example differs from the example in our last article because P_1 and P_2 are both poles, whereas in the last article the corresponding limited portion of the real axis went from a pole to a zero. Now suppose that X is at $-1 - x + j0$ where x is between 0 and 1. Then we have

$$K = R_1\,R_2\,R_3 = x\,(1 - x)\,(4 - x) \quad .. \quad .. \quad (7)$$

and this expression is zero for $x = 0$ and for $x = 1$ and positive for all values of x in between. It therefore has a maximum, which is easily found to be 0·8794 for $x = 0{\cdot}4648$. If therefore K is between 0 and 0·8794, there will be one root between -1 and $-1{\cdot}4648$ and a second between $-1{\cdot}4648$ and -2; as K increases these two roots will approach each other along the real or α-axis until when K reaches the "breakaway value" 0·8794, *the two roots coincide at the "breakaway point"* $-1{\cdot}4648 + j0$. In fact, when K has the "breakaway value",

$$D(p) + K\,N(p) = p^3 + 8p^2 + 17p + 10{\cdot}8794 = (p + 1{\cdot}4648)^2\,(p + 5{\cdot}0704) \ldots \quad (8)$$

and this squared factor in Equ. (8) is not an accident but an asset, of which good use can be made to determine roughly the position of the breakaway point in more difficult cases, as will be explained later. The important point to notice immediately, however, is that if K is increased beyond the "breakaway value", the root-locus leaves the real axis at the breakaway point at right angles. To find what happens to it after that, we can find by trial and error a root for which $\alpha = -1{\cdot}7$ or $-1{\cdot}3$, say, but it will save much vain searching for roots in unpromising areas if we first consider, as in last month's article, what happens when X is a long way off in direction θ, so that $\theta_1 = \theta_2 = \theta_3 = \theta$. To satisfy Equ. (5), θ must either be $\pm 60^\circ$ or 180°. If θ is 180°, this simply means that the corresponding root is real and negative, and that the whole of the real or α-axis to the left of P_1 is part of the root-

locus. As however θ can also be $\pm 60°$, there is an asymptote of the root-locus making $60°$ with the real axis. In our last month's example the asymptote was parallel to the imaginary axis, and we were easily able to find its actual position by taking $1/\omega$-terms into account and neglecting $1/\omega^2$-terms in that case. In our present case it is in fact possible to turn the whole diagram of Fig. 1 anticlockwise through $30°$ and apply a similar argument, but the resulting rule is so simple that it is not worth doing more than quoting it and noting that a proof is given in Reference 1 of last month's article. Asymptotes always go *through the* 'centre of gravity', that is to say, the point determined by the expression

$$G = \frac{(\text{sum of poles}) - (\text{sum of zeros})}{(\text{number of poles}) - (\text{number of zeros})} \quad .. \quad (9)$$

so that in our case $G = -2\frac{2}{3}$. As complex zeros and poles occur in conjugate pairs, G is always on the real axis.

Since there is an asymptote making $60°$ with the real axis, it is clear that for large values of K, there must be roots to the right of the imaginary axis, and therefore there must be a place where the root-locus crosses the imaginary axis, which will be of critical importance in relation to stability. Further, it must be the upper branch of the root-locus that breaks away from the real axis at $-1{\cdot}4648 + j0$, joins this asymptote and crosses the imaginary axis eventually; this branch therefore curves to the right and it is useless to look for a complex root with real part $-1{\cdot}7$ but there will be a complex root with real part $-1{\cdot}3$. Actually, in this particular case, there is no need to use trial and error to find the imaginary part associated with a complex root having given real part, for if $\alpha + j\omega$ is the complex root we are seeking,

$$\begin{aligned} 0 &= \tan(\theta_1 + \theta_2 + \theta_3) \\ &= \frac{\tan\theta_1 + \tan\theta_2 + \tan\theta_3 - \tan\theta_1 \tan\theta_2 \tan\theta_3}{1 - \tan\theta_2 \tan\theta_3 - \tan\theta_3 \tan\theta_1 - \tan\theta_1 \tan\theta_2} \end{aligned} \quad (10)$$

and therefore the numerator of the last member of Equ. (10) is zero, which gives

$$\frac{\omega}{\alpha+1} + \frac{\omega}{\alpha+2} + \frac{\omega}{\alpha+5} = \frac{\omega^3}{(\alpha+1)(\alpha+2)(\alpha+5)} \quad .. \quad (11)$$

so that, as $\omega \neq 0$ for the complex root,

$$\left.\begin{aligned} \omega^2 &= (\alpha+2)(\alpha+5) + (\alpha+5)(\alpha+1) + (\alpha+1(\alpha+2) \\ &= 3\alpha^2 + 16\alpha + 17 \\ &= 3(\alpha+1{\cdot}4648)(\alpha+3{\cdot}86853) \end{aligned}\right\} \quad (12)$$

This checks that $\omega = 0$ at the breakaway point $-1{\cdot}4648 + j0$. The fact that $\omega^2 = 0$ when $\alpha = -3{\cdot}86853$ is irrelevant; although this value genuinely satisfies Equ. (10), the value of $\theta_1 + \theta_2 + \theta_3$ is there not $180°$ but zero. In particular Equ. (12) shows that $\omega^2 = 17$ when $\alpha = 0$, and we reach the borderline of instability. From Equ. (6), the corresponding value of K is $(18 \times 21 \times 42)^{\frac{1}{2}} = 126$, and therefore the maximum value of $A_0\beta$ is $K/10$ or $12{\cdot}6$. This checks with the fact that when Equ. (3) is cleared of fractions, it becomes

$$p^3 + 8p^2 + 17p + (10 + K) = 0 \quad .. \quad .. \quad (13)$$

and it is well known that for critical stability (that is, a purely imaginary root) in the case of a cubic equation, the condition is that the product of the coefficients of p and p^2 must equal the product of the coefficient of p^3 and the number term. This confirms that the critical value of K is 126 in the case of Equ. (13), and shows that in the more general case of the equation $1 + A\beta = 0$ with $A\beta$ given by Equ. (2), the critical value of $A_0\beta$ is

$$\begin{aligned} &T_2/T_3 + T_3/T_2 + T_3/T_1 \\ &\quad + T_1/T_3 + T_1/T_2 + T_2/T_1 + 2 \quad .. \quad .. \quad (14) \end{aligned}$$

so that, if the main consideration is to make the expression (14) as large as possible, T_1 and T_2 should be made nearly equal and T_3 should be very different—many times larger or smaller than T_1 and T_2.

As Equ. (13) is only a cubic, we can easily obtain a formula for the value of K when we know the real part α of a root; when this formula is combined with Equ. (12) it means that in this particular case (or indeed any case where the roots are given by a cubic equation) we do not in fact need to use trial and error to determine the root-locus or any part of it. But the root-locus trial-and-error technique is available whatever the degree of the equation giving the root whereas a formula for the value of K in terms of α rapidly becomes more difficult to obtain as the degree of the equation rises.

If in Equ. (13), there are two conjugate complex roots $\alpha + j\omega$ and $\alpha - j\omega$, the third root is $-8 - 2\alpha$, since the sum of the three roots must be -8, that is, minus the ratio of the coefficient of p^2 to that of p^3. Hence the factors of the left-hand side of Equ. (13) must be

$$[p^2 - 2\alpha p + (\alpha^2 + \omega^2)]\,[p + 8 + 2\alpha] \quad .. \quad (15)$$

It follows, by equating coefficients of p and the constant terms, that

$$\alpha^2 + \omega^2 - 2\alpha(8 + 2\alpha) = 17 \quad .. \quad .. \quad .. \quad (16)$$

$$\alpha^2 + \omega^2 = (10 + K)/(8 + 2\alpha) \quad .. \quad .. \quad .. \quad (17)$$

and, substituting for ω^2 from Equ. (17) into Equ. (16) and rearranging, we thus find

$$K = 126 + 162\alpha + 64\alpha^2 + 8\alpha^3 \quad .. \quad .. \quad (18)$$

A rough diagram of the root-locus is given in Fig. 2.

It now remains to express the ideas we have used to determine the positions of key portions of the root-locus in a more general form, so that the locus can be determined, when we have not the comforting, confirmatory help of explicit formulae like Equs (12) and (18), with as little trial-and-error determination of complex roots as possible.

First, if Equ. (1) is cleared of fractions and K is zero, it is obvious that the roots are the same as the poles, so the root-locus consists of a number of branches which begin at the poles.

Second, in both examples we found that certain portions of the real or α-axis belonged to the root-locus while the remainder did not. The general rule here is that a point of the real axis belongs to the locus if and only if there is an odd number of real poles and/or zeros to the right of it. Complex poles or zeros, which occur in conjugate pairs, are irrelevant in this connection.

Next, there is the question of the position and direction of asymptotes to branches that go to infinity. The

rules here are quite simple: asymptotic directions are obtained by dividing any odd multiple of 180° by the difference between the number of poles and the number of zeros. In last month's example, this difference was 1, so there was only one asymptote at 180°, and when this happens the asymptote is always the real axis itself. But in this month's example (see Fig. 2) there were three poles and no zeros, so asymptotic directions occur at odd multiples of 60°. The rule for finding the position of G, the centre of gravity of the roots, where the asymptotes meet the real axis, has already been given in Equ. (9).

Next, we have to consider how to determine the breakaway point roughly and easily. Now we noticed that in Fig. 2 and Equ. (8), K had a maximum value (for real roots) at the breakaway point. It follows that log K also has a maximum and therefore its derivative is zero. [This is equally true in cases where a portion of the real axis belonging to the root-locus joins two zeros instead of two poles as in Fig. 2; in such cases K has a minimum (for real roots) at the breakaway point between those zeros.] Now it can be shown that

$$\frac{d}{dp}\log K = \frac{1}{p-p_1} + \frac{1}{p-p_2} + \frac{1}{p-p_3} + \ldots - \frac{1}{p-z_1} - \frac{1}{p-z_2} - \frac{1}{p-z_3} - \ldots \quad (19)$$

where p_1, p_2, p_3 etc. are the various poles (real or complex) and z_1, z_2, z_3 etc. are the various zeros. Hence the right-hand side of Equ. (19) can be equated to zero if p_b, corresponding to the breakaway point, is substituted for p. The real poles and zeros present no difficulty in Equ. (19) except insofar as care must be taken to get the signs right. Complex poles (and zeros) must be taken together in conjugate pairs; if for example $p_2 = \alpha + j\omega$ and $p_3 = \alpha - j\omega$, we have

$$\frac{1}{p_b-\alpha-j\omega} + \frac{1}{p_b-\alpha+j\omega} = \frac{2(p_b-\alpha)}{(p_b-\alpha)^2+\omega^2} = \frac{2(p_b-\alpha)}{R^2} \quad .. \quad (20)$$

R being the distance from p_b to $\alpha \pm j\omega$, which is easily measured.

It is not usually necessary to determine the breakaway point very accurately, and if in Equ. (20) ω is large, the contribution from the two complex conjugate poles is small. A good idea of the position of the breakaway point can therefore often be obtained graphically by applying Equ. (19) with only real or nearly real poles and zeros taken into account. The determination of K for any *real* value of p is not difficult, so if greater accuracy is required, the variation of K for real values of p near the value obtained graphically can be checked to see that there really is a maximum or minimum in that neighbourhood, but usually in Equ. (19) one or two terms on the right-hand side are much more sensitive to small variations of p near the suspected breakaway point than the remainder, and Equ. (19) thus gives p_b easily.

The only remaining points that help in drawing the root-locus are, that at the breakaway point the root-locus always leaves the real axis at right angles, and that the direction in which it leaves complex poles (or reaches complex zeros) can be determined, as in last month's example, by finding the phase angles of all the lines joining the particular pole (or zero) to the remaining poles and zeros, and thus determining the angle of departure or arrival which will make the phase of $A\beta$ equal to an odd multiple of 180°.

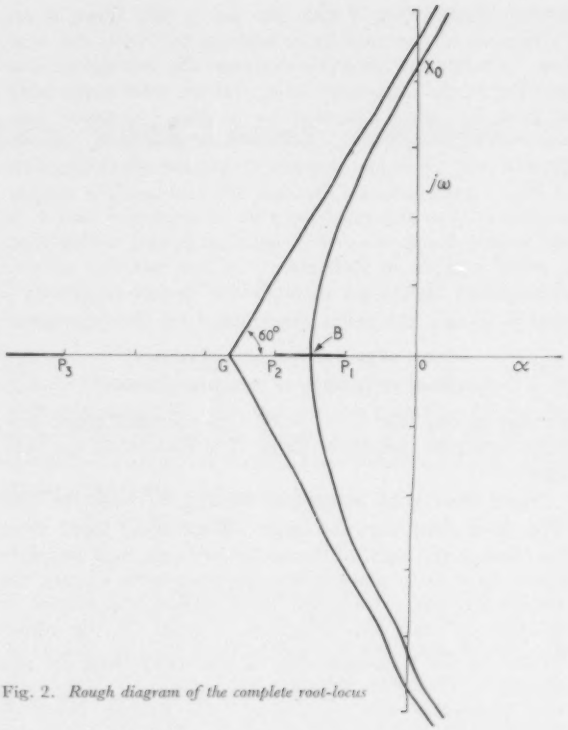


Fig. 2. *Rough diagram of the complete root-locus*

It will be seen that all the above rules have been applied in the two examples given: Equ. (19) tells us that in the particular example discussed here, the breakaway point is given by

$$\frac{1}{p_b+1} + \frac{1}{p_b+2} + \frac{1}{p_b+5} = 0 \quad .. \quad .. \quad (21)$$

and this agrees with Equs (11) and (12), although in this case it was quite easy to obtain the breakaway point directly from the fact that K was a maximum there. The application of the remaining rules to the two examples given is straightforward.

ALMA ACQUIRE ALSTON CAPACITORS

Alma Components Ltd. have announced that they have purchased the whole of the issued share capital of Alston Capacitors Ltd., of Halesworth, Suffolk. Under the new ownership Alston will continue to manufacture Type Approved capacitors, including silver mica Types GMMIG and GMM3Y.

INTERNATIONAL ELECTRONIC COMPONENTS EXHIBITION—PARIS 1962

The Fédération Nationale des Industries Electroniques have announced that the International Exhibition of Electronic Components will be held from 16th to 20th February 1962 in Paris. Further detail are available from Mr. R. Foucault, 23 rue de Lübeck, Paris 16e.

Finite-Gap Klystrons

A LARGE-SIGNAL BALLISTIC THEORY

By L. Solymar*

SUMMARY. *The amplitudes of the fundamental and some higher-harmonic components of current are calculated and plotted for a number of different gap widths and modulations for finite-gap klystrons. It is shown that as the a.c. voltage becomes comparable with the d.c. voltage the dependence on gap width is considerably altered but the gap admittance is practically invariant.*

More than twenty years ago the first klystrons and the theory describing them were produced, but still a complete mathematical explanation of their operation seems remote. The underlying physics involved is clear enough, yet all the theoretical approaches so far tried involve drastic simplifying assumptions and consequently major uncertainties in the validity of some of the conclusions. This is particularly true with regard to predictions relating to high-efficiency operation in fundamental or in frequency-multiplier klystrons.

The small-signal solution for both infinite and finite beams is well known and, in most cases, is capable of predicting experimental results with considerable accuracy. Attempts at devising a large-signal theory (analytic or numerical) have been reported only in fairly recent times. Two radically different approaches have been described: in the first, due to Paschke[1], the starting point is the small-signal space-charge wave theory which is extended by the inclusion of non-linear terms. On the other side, there are approaches such as the 'disc model' introduced by Webber[2], in which the starting point is ballistic theory, modified by the presence of space charge.

Webber has obtained a wide range of solutions using computer techniques. In the course of these he had to make a number of simplifying assumptions; in particular he confined his investigation to the case of an infinitesimally-short modulating gap, and sinusoidal velocity modulation. The present work consists of a numerical study of purely ballistic theory (space charge neglected), in which these two restrictions are removed.

The Movement of the Electrons in the Gap

Assuming a homogeneous electric field in the gap and neglecting space-charge effects, the Newtonian equation for the movement of an electron can be written as follows:

$$\frac{d^2z}{dt^2} = -\frac{e}{m}\frac{V}{d}\sin\omega t \quad .. \quad .. \quad .. \quad .. \quad (1)$$

where

z = the co-ordinate in the direction of the electron movement (Fig. 1).

e = the charge of an electron.
m = the mass of an electron.
V = a.c. voltage.
ω = the angular frequency.
d = the width of the gap.

Introducing the notations

$$\zeta = \frac{\omega z}{u_0}, \quad D = \frac{\omega d}{u_0}, \quad \phi = \omega t$$

$$\alpha = \frac{V}{V_0}, \quad \gamma = \frac{\alpha}{2D}, \quad \mu = \frac{u}{u_0} \quad .. \quad .. \quad .. \quad (2)$$

Equ. (1) can be transformed into the following form[3]:

$$\frac{d^2\zeta}{d\phi^2} = \gamma \sin\phi \quad .. \quad .. \quad .. \quad .. \quad .. \quad (3)$$

where V_0 = d.c. voltage, u_0 = d.c. velocity, and u = a.c. velocity.

Applying the boundary conditions ($\phi = \phi_0$ and $d\zeta/d\phi = 1$ at $\zeta = 0$) the solution can be given as follows:

$$\mu = 1 + \gamma(\cos\phi_0 - \cos\phi) \quad .. \quad .. \quad .. \quad (4)$$

and

$$\zeta = (1 + \gamma\cos\phi_0)(\phi - \phi_0) - \gamma(\sin\phi - \sin\phi_0) \quad (5)$$

The Current at an Arbitrary Point in the Drift Tube

According to the principle of the conservation of charge the currents at the input of the gap and at an

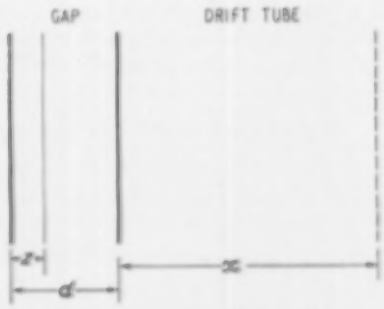
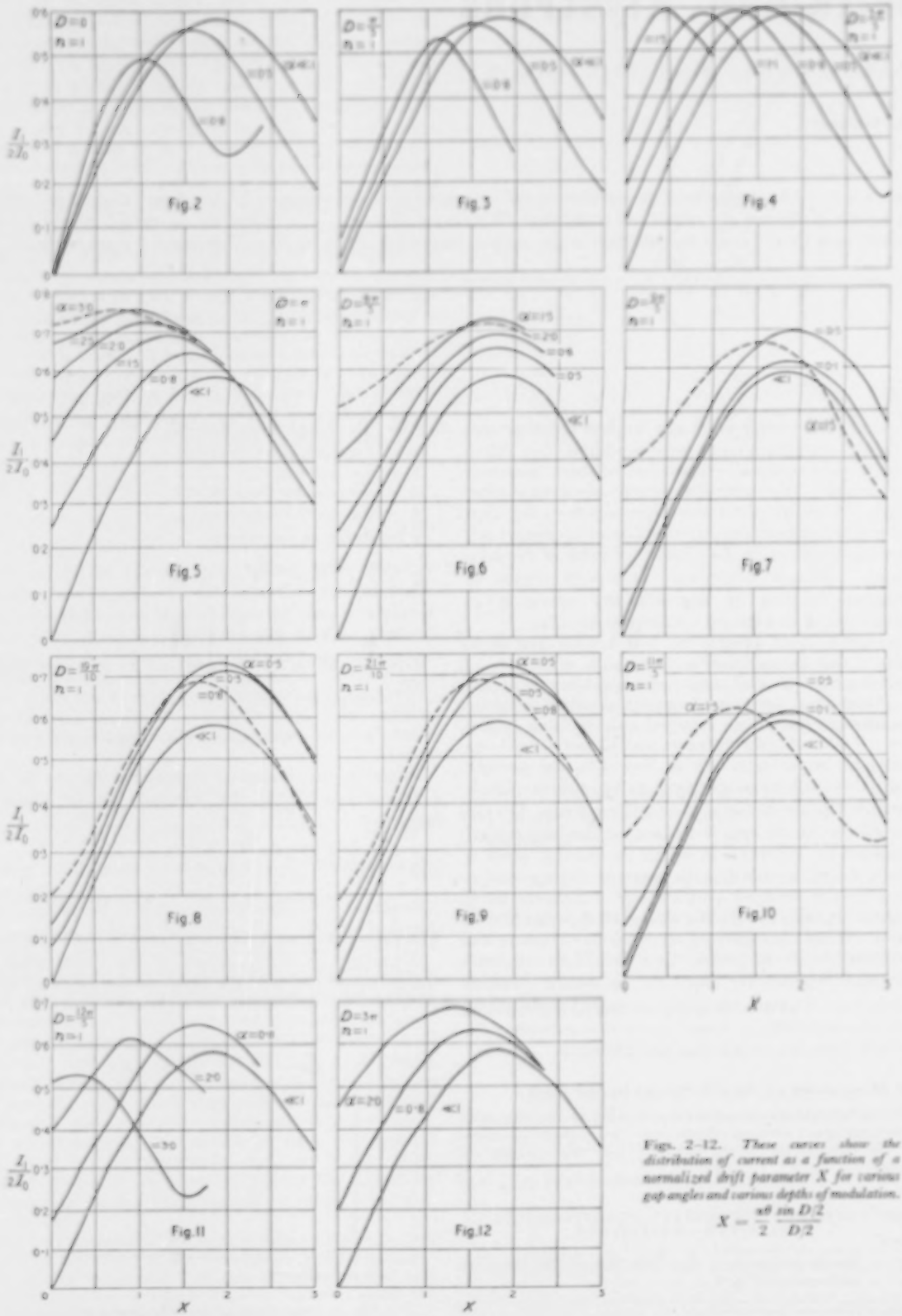


Fig. 1. *Simple diagram showing the dimensions z, d and x*

* *Standard Telecommunication Laboratories Ltd., Harlow.*



Figs. 2–12. *These curves show the distribution of current as a function of a normalized drift parameter X for various gap angles and various depths of modulation.*

$$X = \frac{\omega\theta}{2} \frac{\sin D/2}{D/2}$$

arbitrary point in the drift tube are related by the simple formula

$$I = I_0 \frac{d\phi_0}{d\phi} \qquad (6)$$

The relation between ϕ and ϕ_0 can be simply calculated from the movement and velocity of the electrons

$$\phi = \phi_s + \frac{\theta}{\mu(\phi_s)} = \phi_s + \frac{\theta}{1 + \gamma(\cos\phi_0 - \cos\phi_s)} \qquad (7)$$

where ϕ_s corresponds to the time at which the electrons leave the gap, $\mu(\phi_s)$ is the output velocity, and $\theta = x/u_0$ is the normalized drift length.

The nth harmonic of the current can be obtained by expanding I into a Fourier series

$$I_n = \frac{1}{\pi}\int_{-\pi}^{\pi} I \exp(jn\phi)\, d\phi$$

$$= \frac{I_0}{\pi}\int_{-\pi}^{\pi} \exp\left[jn\left[\phi_s + \frac{\theta}{1+\gamma(\cos\phi_0 - \cos\phi_s)}\right]\right] d\phi_0 \qquad (8)$$

where ϕ_0 and ϕ_s are connected by the implicit equation

$$D = (1+\gamma\cos\phi_0)(\phi_s - \phi_0) - \gamma(\sin\phi_s - \sin\phi_0) \qquad (9)$$

The integration was carried out with the aid of the Stantec 'Zebra' electronic computer. The results are discussed in the following sections.

The Fundamental Component of Current

As most klystrons are operating on the fundamental component of current, this quantity is of primary interest. Accordingly, $I_1/2I_0$ is plotted in Figs. 2 to 12 for different values of x and D as a function of

$$X = \frac{x\theta}{2}\frac{\sin D/2}{D/2} \qquad (10)$$

This choice of the abscissa enables us to compare the results with those of the small-signal ballistic theory, where the fundamental current is given by the first-order Bessel function.

For the design of a high-efficiency klystron the position and the magnitude of the current maximum is of particular importance. As can be seen in Figs. 2 to 12 these two parameters depend very much on the gap angle.

For zero gap angle (Fig. 2) the higher the modulation the smaller is the available current maximum. As the gap angle is increasing this tendency is inverted. At $D = 0{\cdot}6\,\pi$ (Fig. 4) the available current maxima are roughly independent of x. At $D = \pi$ (Fig. 5) the higher the modulation the larger is the current maximum reaching 0·755 at $x = 2{\cdot}5$. At $D = 1{\cdot}6\,\pi$, $1{\cdot}8\,\pi$, $1{\cdot}9\,\pi$, (Figs. 6 to 8) higher modulation results in higher current maxima but only up to a certain value of x. If x exceeds that limit the current maxima begin to decrease again. The region between $2\,\pi$ and $3\,\pi$ (Figs. 9 to 12) exhibits roughly the same properties as that between $2\,\pi$ and π.

It is interesting to note that for gap angles near to $2\,\pi$ even small values of x can result in a large current maximum. This significant increase can be attributed to the effect of the second-harmonic velocity excitation as approximately calculated by Feenberg[3] for small signals.

The optimum-drift distance is determined by the value of X_m where the current maximum occurs. It can be seen that except the region near to $2\,\pi$ higher modulation results in smaller X_m.

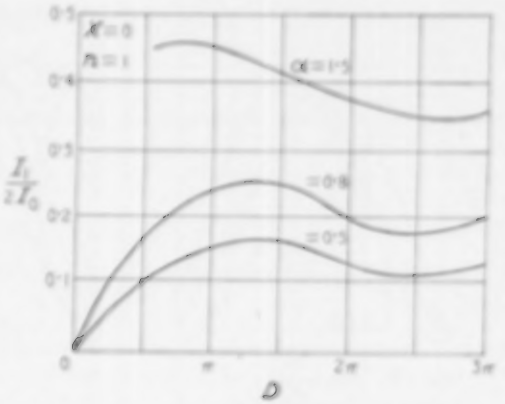


Fig. 13. *By showing $I_1/2I_0$ (at $X = 0$) plotted against D, this clearly reveals that for higher values of x and D the fundamental current at the input of the drift tube is already appreciable*

A look at Figs. 2 to 12 reveals that for higher values of x and D the fundamental current at the input of the drift tube is already appreciable. To show this relationship more clearly the value of $I_1/2I_0$ at $X = 0$ is plotted in Fig. 13 against D.

Higher Harmonics of the Current

Similar calculations have been made for $n = 2$, for a number of different gap angles and modulations, which are shown in Figs. 14 to 22. The numerical values of the positions and magnitudes of the current maxima are obviously different, but the qualitative dependence on the gap angle is unchanged.

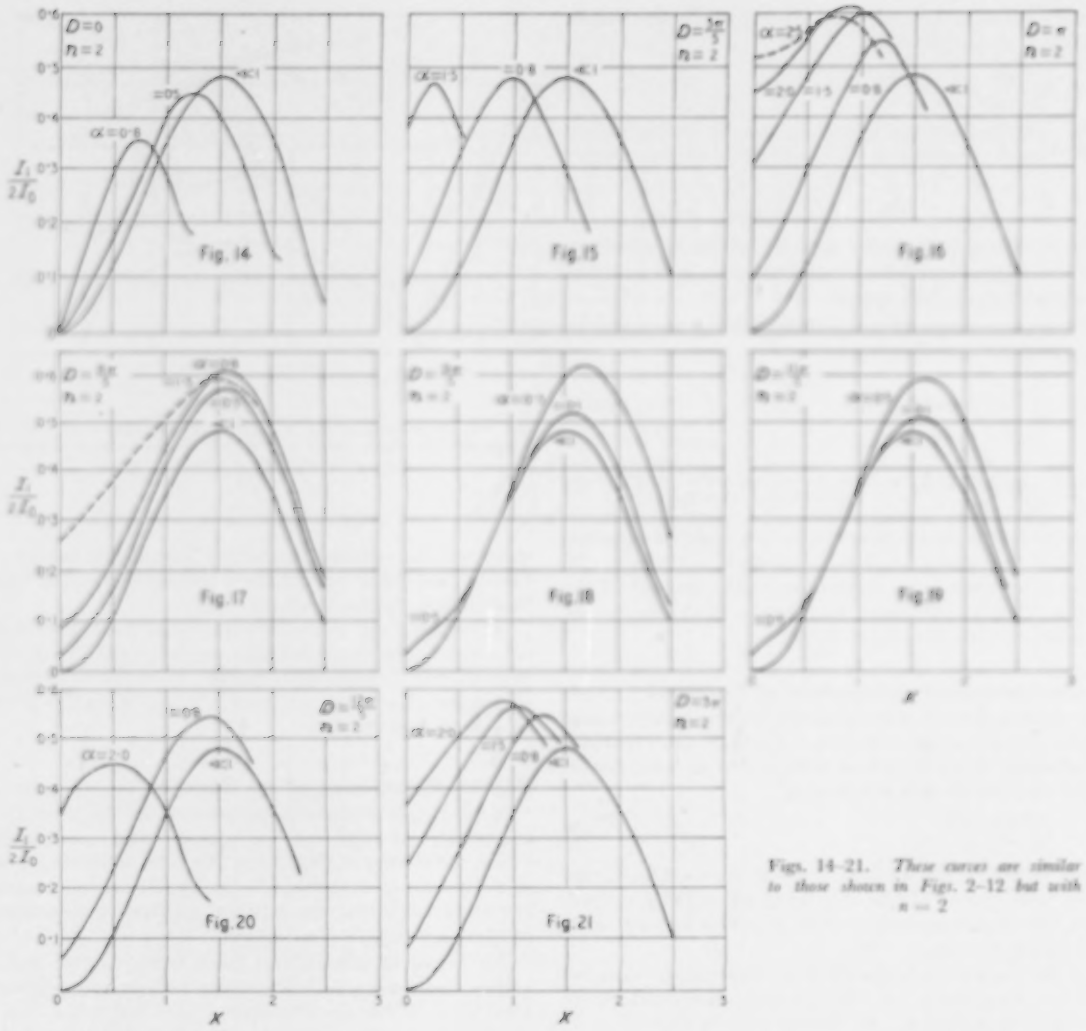
No detailed calculations have been carried out for higher harmonics. A few special cases are plotted in Figs. 23 to 25 for $x = 0{\cdot}5$ and $D = 1{\cdot}6\,\pi$, $1{\cdot}8\,\pi$, $2\,\pi$, for the first four harmonics. It can be seen that as the gap angle increases from $1{\cdot}6\,\pi$ to $1{\cdot}8\,\pi$ both the optimum drift distances and the available current maxima increase unanimously for all the four harmonics. This would suggest that the conclusions are qualitatively valid for higher harmonics as well.

At $D = 2\,\pi$ the gap factor is zero and according to the small-signal theory there is no velocity modulation. However, for large signals there is a small velocity modulation containing mainly even harmonics. Thus —as shown in Fig. 25—the current for the even harmonics is considerably larger than that for the odd ones.

Calculation of the Beam Admittance

To obtain the loading effect of the beam in the buncher gap the average current has first to be computed. The fundamental component of the average current can be written as follows[4]

$$I_1 = \frac{1}{\pi}\int_{-\pi}^{\pi} \exp(-j\phi)\left[\frac{1}{d}\int_0^d I dz\right] d\phi \qquad (11)$$



Figs. 14–21. *These curves are similar to those shown in Figs. 2–12 but with* $n = 2$

Fig. 22. *These curves are analogous to those of Fig.* 13 *but for* $n = 2$

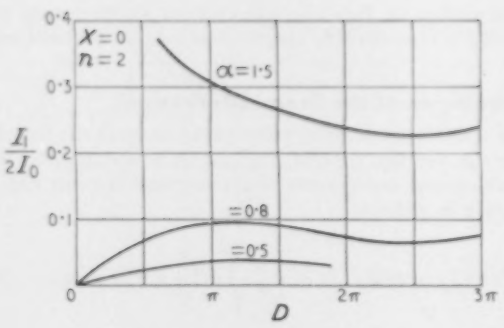
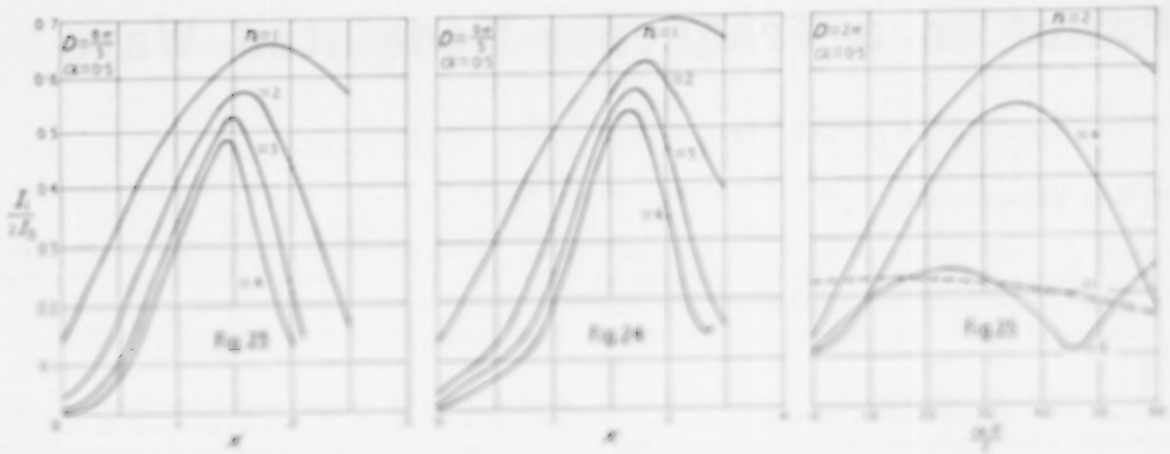


TABLE 1

α \ D	π	1·4π	1·8π	2·2π	2·6π	3π
≪1	0·202	0·175	0·058	−0·038	−0·038	0·022
	0·000	−0·128	−0·178	−0·118	−0·028	0·000
0·8	0·206	0·172	0·056	−0·038	−0·037	0·023
	−0·007	−0·130	−0·177	−0·117	−0·028	−0·001
1·5	0·212	0·164	0·053	−0·036	−0·035	0·023
	−0·028	−0·136	−0·172	−0·115	−0·028	−0·004

G_B/G_0 (first row) and B_B/G_0 (second row) for different values of α (modulation) and D (gap angle) where $G_0 = I_0/V_0$.



Figs. 23-25. *A few special cases for* $\alpha = 0{\cdot}5$ *and* $D = 1{\cdot}6\,\pi$, $1{\cdot}8\,\pi$, *and* $2\,\pi$ *for the first four harmonics.*

Performing the operations in Eqs. (11) we get the admittance

$$\frac{Yf}{Y_0} = j\,\frac{1}{2\pi^2 \alpha}\int_{-\pi}^{\pi} (\phi_a - \phi)\,(\cos\phi_a + \cos\phi) + 2\,(\sin\phi - \sin\phi_a)\,e^{-jn\phi}\,d\phi \quad \ldots \quad (12)$$

where the implicit relationship between ϕ and ϕ_a is given again by Eqs. (9) replacing ϕ_a by ϕ. Carrying out the integration numerically Table 1 shows the real and imaginary part of the beam admittance for different values of α and D. It can be seen that excitation by large signals makes practically no difference.

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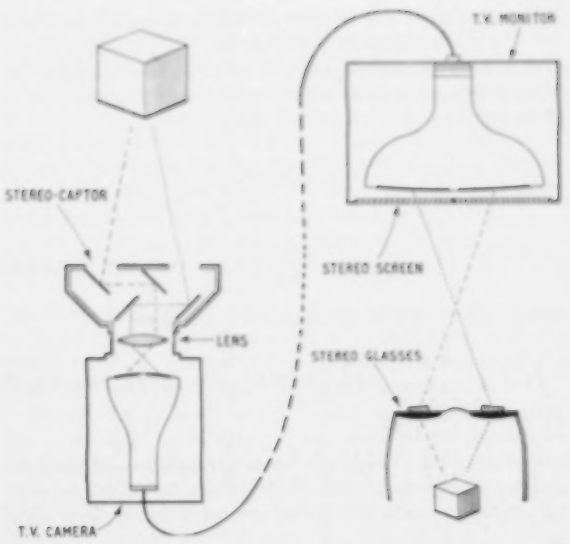
THREE-DIMENSIONAL TELEVISION

A stereoscopic television kit which is designed to convert any closed-circuit television to a three-dimensional system, with no electronic modifications, has been announced by Stereotronics Corporation, U.S.A.

This is an optical kit and consists of a 'Stereo-Captor' which fits on the television camera lens, a 'Stereo-Screen' to fit in front of the receiving monitor tube, and 'Stereo-Glasses'.



The picture illustrates the component parts of the optical system while the schematic diagram indicates the principle of operation.

Power in a Frequency-Modulated Wave

DURING A FINITE INTERVAL

By R. G. Medhurst*, B.Sc.

This article is prompted by a recent revival of interest[1,2] in the evaluation of the mean power associated with a frequency-modulated wave, both long-term and over a finite period. A well-known result in the theory of frequency modulation is that the sum of the mean long-term powers in all the sidebands, together with the residual power in the carrier, is equal to the mean long-term power in the unmodulated carrier. In Ref. 1, this result is developed (for single-tone modulation) from two lines of argument, neither of which is claimed to be entirely satisfactory. Some years ago, Kulp[3] produced what appears to be a convincing demonstration of the theorem in question, and he was able to extend his proof to cover modulation by an arbitrary number of tones.

In Ref. 2, the argument of Ref. 1 is modified, and further consideration given to the mean power of the modulated wave measured over a finite period. The analysis is restricted to a sampling time which is exactly equal to one period of the modulating wave (single-tone modulation being assumed). A numerical result is obtained only when the carrier frequency is an exact multiple of the modulation frequency. In this case, it is shown that the mean power over this particular sampling period is independent of the initial time at which the sample is taken. When the carrier frequency is not an exact multiple of the modulation frequency, the mean power in the sampling period considered in Refs. 1 and 2 will depend on the initial instant of the sample. It is suggested in Ref. 2 that no simple expression is available for the resulting variation with time of the mean power. The purpose of the present article is to show that a good approximation to such an expression is obtainable quite easily. Furthermore, there is no difficulty in extending the result to cover sampling times not equal to the period of the modulation. An estimate given in Ref. 2 for the variation with time of the mean power in one modulation period is shown to be very much too small.

Using the notation of Ref. 1, the instantaneous power delivered by the modulated wave to a resistance R is

$$\frac{V_c^2}{2R}[1 + \cos(2\omega_c t - 2m \sin \omega_m t)] \quad .. \quad .. \quad (1)$$

where V_c is the peak voltage, ω_c the angular carrier frequency, ω_m the angular modulation frequency and m the modulation index. It is to be noted that $V_c^2/2R$ is the long-term average power in the unmodulated carrier. The average power measured over one modulation period, commencing at time T, is

$$\frac{V_c^2}{2R}[1 + A] \quad .. \quad .. \quad .. \quad .. \quad .. \quad (2)$$

where

$$A = \frac{\omega_m}{2\pi}\int_T^{T+\frac{2\pi}{\omega_m}} \cos(2\omega_c t - 2m \sin \omega_m t)dt \quad .. \quad (3)$$

Putting $\omega_m T = \alpha$, $2\omega_c/\omega_m = k$, $2m = p$,

$$A = \frac{1}{2\pi}\int_\alpha^{\alpha+2\pi} \cos(kx - p \sin x)dx \quad .. \quad (4)$$

When k is an integer, $=n$, say, Equ. (4) is recognizable as defining a Bessel coefficient of the first kind, and

$$A = J_n(p), \quad .. \quad .. \quad .. \quad .. \quad .. \quad (5)$$

independent of α, which is Vaughan's result. When k is not an integer, A becomes dependent on α. For $\alpha = 0$, Equ. (4) defines the original form of Anger's function[4], and, unlike the Bessel coefficients, does not appear to have been tabulated to any useful degree. The general form when α is not zero, is still less a tabulated function. Thus, one seeks an approximation, of adequate accuracy.

Suppose that $k = n + \delta$, where n is an integer and $|\delta| \leqslant \frac{1}{2}$. In normal practice, n will be quite large, so that it is natural to consider an expansion of Equ. (4) in inverse powers of n. In order to retain the value of A, given by Equ. (5), when δ is zero, write A as

$$\frac{1}{2\pi}\int_\alpha^{\alpha+2\pi} \cos(nx - p \sin x)dx + \frac{1}{2\pi}\int_\alpha^{\alpha+2\pi} [\cos(nx + \delta x - p \sin x) - \cos(nx - p \sin x)]dx$$

which, after some re-arrangement, becomes

$$J_n(p) - \frac{1}{\pi}\int_\alpha^{\alpha+2\pi} \sin\left(\frac{\delta}{2}x\right)\sin\left(\frac{\delta}{2}x - p \sin x\right)\cos(nx)\,dx - \frac{1}{\pi}\int_\alpha^{\alpha+2\pi} \sin\left(\frac{\delta}{2}x\right)\cos\left(\frac{\delta}{2}x - p \sin x\right)\sin(nx)\,dx \quad .. \quad (6)$$

*The General Electric Company Limited, Telecommunications and Engineering Research Laboratories, Hirst Research Centre, Wembley, England.

Expression of the form $\int_{\alpha}^{\alpha+2\pi} f(x) \begin{matrix}\sin\\ \cos\end{matrix} (nx)\ dx$ can immediately be expanded in inverse powers of n by successive integrations by parts. In fact,

$$\int_{\alpha}^{\alpha+2\pi} f(x)\sin(nx)dx = -\frac{1}{n}[f(\alpha+2\pi)-f(\alpha)]\cos(n\alpha) + \frac{1}{n^2}[f'(\alpha+2\pi)-f'(\alpha)]\sin(n\alpha) + - - - - - \quad .. \quad (7)$$

and

$$\int_{\alpha}^{\alpha+2\pi} f(x)\cos nx\,dx = \frac{1}{n}[f(\alpha+2\pi)-f(\alpha)]\sin(n\alpha) + \frac{1}{n^2}[f'(\alpha+2\pi)-f'(\alpha)]\cos(n\alpha) - - - - - - .. \quad .. \quad (8)$$

Using Equs (7) and (8), the required expansion of Equ. (6) readily follows. For any case likely to arise in practice, it is probably sufficient to work to order $1/n$ only, though more terms could be found without difficulty. Then, to order $1/n$, after some re-arrangement, it is found that

$$A = J_n(p) + \frac{1}{\pi n}\sin(\pi\delta)\cos[n\alpha+(\alpha+\pi)\delta-p\sin\alpha] \quad .. \quad (9)$$

Putting this back into the original notation, we find that the mean power measured in a time equal to one modulation period is, approximately,

$$\frac{V_c^2}{2R}\left\{1+J_{2\omega_c/\omega_m}(2m) + \frac{1}{2\pi}\frac{\omega_m}{\omega_c}\sin(\pi\delta)\cos[2\omega_c T-2m\sin(\omega_m T)+\pi\delta]\right\} \quad .. \quad (10)$$

When $\delta = 0$, this becomes independent of T, the time of commencement of the sampling, as previously shown by Vaughan. When $\delta \neq 0$, the measured mean power in this sampling period would depend on the instant of sampling, the spread of the samples being given approximately by the factor $[1 \pm \omega_m/(2\pi\omega_c)]$. In Ref. 2, a tentative suggestion for this factor is $[1 \pm J_{2\omega_c/\omega_m}(2m)]$. This is, in fact, too small by many orders, since, for large values of $2\omega_c/\omega_m$ and practical values of m,

$$J_{2\omega_c/\omega_m}(2m) \simeq \frac{1}{\sqrt{2\pi}}\left(\frac{2\omega_c}{\omega_m}\right)^{-2\omega_c/\omega_m} \quad .. \quad .. \quad (11)$$

(see Ref. 4, p. 225). However, in any situation likely to arise in radio engineering, it is still true that the fluctuation of the power requirement indicated by Equ. (10) will be of quite negligible importance.

It may be as well to point out, in conclusion, that the constancy with time of sampled mean power for integral values of $2\omega_c/\omega_m$ only holds when the sampling time is an exact multiple of one modulation period. This may be seen by repeating the above analysis for a general sampling time T'. It is found, after carrying out the algebra, that, to order $\omega_m/2\omega_c$, A [defined by Equ. (2)] is given by

$$A = \frac{1}{\omega_c T'}\sin\left\{\omega_c T' - 2m\sin(\tfrac{1}{2}\omega_m T')\cos(\omega_m T+\tfrac{1}{2}\omega_m T')\right\}\cos\left\{2\omega_c T+\omega_c T' - 2m\cos(\tfrac{1}{2}\omega_m T')\sin(\omega_m T+\tfrac{1}{2}\omega_m T')\right\} \quad .. \quad .. \quad .. \quad (12)$$

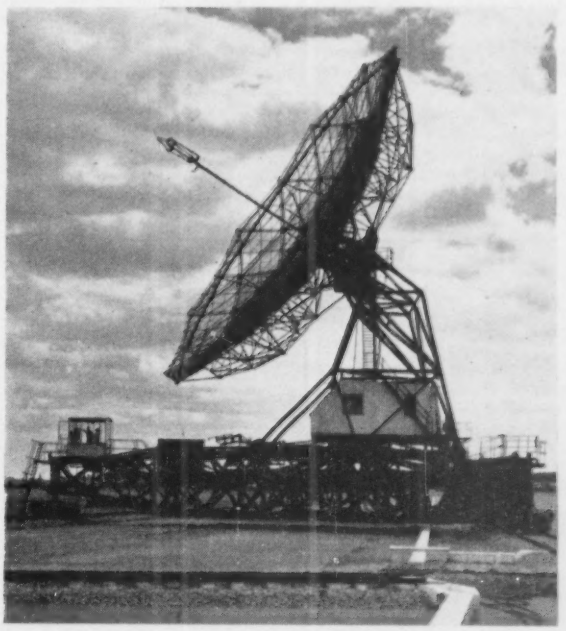
This corresponds to the third term in the curly bracket of Equ. (10). It is clear that for A to be independent of T, $\frac{1}{2}\omega_m T'$ must be an exact multiple of π, in addition to the requirement that $\omega_c T'$ should be an exact multiple of π.

REFERENCES

[1] W. C. Vaughan, "Power in an Angle-Modulated Wave", *Electronic Radio Engr.*, Aug. 1959, Vol. 36, pp. 289–291.
[2] W. C. Vaughan, "Angle-Modulated Wave: A Further Note on Associated Power", *Electronic Technol.*, Oct. 1960, Vol. 37, pp. 387–388.
[3] M. Kulp, "Über der Effektivwert Modulierter Schwingungen", *Frequenz*, Oct. 1952, Vol. 6, pp. 290–295.
[4] G. N. Watson, "A Treatise on the Theory of Bessel Functions", Cambridge University Press, 2nd Edition, 1952, p. 308.

R.R.E INTERFEROMETER

This picture shows one of the two 82-ft. diam. steerable and mobile radio telescopes which have been installed for radio astronomical and upper atmosphere research at the Royal Radar Establishment at Defford near Malvern. Each telescope is mounted on its own railway track, and the two are laid along intersecting runways in the form of a letter 'T', the length of the tracks from the



intersection being 750 metres. As well as being used together as an interferometer, the two radio telescopes can also be used separately. The pointing accuracy of each 25-metre reflector, which give beam-widths of about 20 minutes of arc at $\lambda = 10$ cm, should not exceed 2 to 3 minutes of arc. The angular accuracy of the interferometer, with a baseline spacing between the two telescopes of 1000 metres, is a few seconds of arc.

Correspondence

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Zener Diode in Stabilized Transistor Power Supplies

SIR,—In a letter in the June issue, Messrs. Dutta Gupta and Tyler give a criterion for the practicability of a stabilizer in the form of an inequality which must be satisfied. They have omitted to indicate the factor by which the inequality must be exceeded to make the stabilizer not only practicable but useful. If the ratio between the maximum permissible diode current (i_{Zx}) and the maximum current extracted from it by the load (i_{Lx}) is γ, then their inequality, equation (10), becomes

$$\gamma > 1 \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (10a)$$

and (11) becomes

$$\gamma > 1+\lambda \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (11a)$$

where λ is the ratio of the input voltage range ($V_{Ix} - V_{I\beta}$) to the minimum voltage ($V_{I\beta}$). The analysis can then be extended to show that the Zener diode voltage (V_Z) is related to the minimum input voltage ($V_{I\beta}$) by

$$\frac{V_Z}{V_{I\beta}} = 1 - \frac{\lambda}{\gamma - 1} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (13)$$

If we now put $\gamma > 1+\lambda$ in (13), it is clear that the criterion for (11) is simply $V_Z > 0$.

As an example of a circuit which could easily satisfy this criterion, the authors give Fig. 4 in which the value of the constant is $\gamma = 900/375 = 2{\cdot}4$. In this case, $V_Z/V_{I\beta} = 0{\cdot}64$ so that, in stabilizing an input voltage varying between, say, 6 V and 9 V, the output voltage can be no more than 3·9 V. This inefficiency seems to be a result of their choice of a diode with a maximum current of 8 mA and one wonders why a more generously-rated diode was not used.

It is often of great advantage to produce a stable output voltage close to the minimum input voltage particularly, for example, in decoupling the supply to an oscillator from the supplies to the rest of a receiver. A circuit which has proved very useful in this respect is shown in Fig. A, which is similar (apart from polarity) to the Fig. 1 of Messrs. Dutta Gupta and Tyler except that the resistor R is replaced by a transistor, forming an almost constant-current device feeding a constant-voltage device.

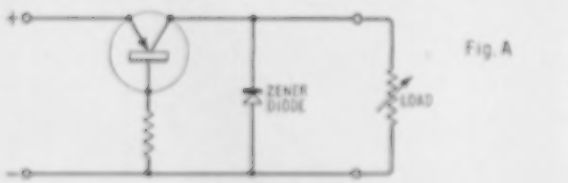


Fig. A

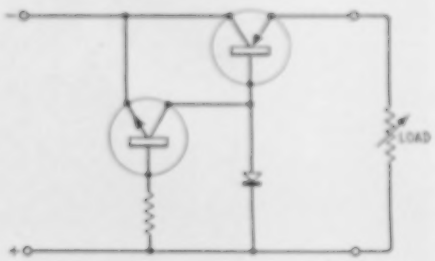


Fig. B

This circuit is inefficient in that the effective load on the source is a constant resistance but this is of little consequence when the circuit is substituted in Figs. 3 and 4. Fig. 3 then becomes as Fig. B, the change of polarity being only to avoid the expense of an n-p-n power transistor.

The advantage of this modification is that the restriction of (13) no longer applies and V_Z can take any value up to $V_{I\beta} - V_{Ix}$ provided only that $\gamma > 1+\lambda$.

R. V. HARVEY.

Research Department,
British Broadcasting Corporation,
Kingswood, Surrey.
27th July 1961.

Transistor Circuit Diagrams

SIR,—Mr. J. C. S. Richards (August 1961, p. 306) is a teacher dealing with malleable minds of graduate calibre; in these circumstances there should be no difficulty in communication, whatever the approach, and I agree that the logic of conventional current flow may be appropriate when dealing with intelligent, but inexperienced, students. However, I am charged with responsibility for the design and engineering of equipment for small and medium batch production in a highly competitive field and, with previous direct responsibility for rather more than £4 × 10⁶ of such equipment, I believe that my views have some relevance.

The disposition of the active element within a circuit diagram is a poor peg upon which to hang an advancing technology; it must be assumed that engineers will accept any reasonable convention, as conventional current flow has been accepted. In fact, the active device may, almost, be considered an irrelevance during the initial design stage when working with the equivalent circuit, and I have yet to see an equivalent circuit with the elements drawn below the reference line. When the initial design has been completed it now has to be realized as a practical proposition by a worker who may be a non-graduate apprentice or who may be a well-experienced engineer. After satisfactory development a complete equipment goes through the factory and into the field. At all stages, after design, information has to be communicated to workers of varying skills who have, however, all had previous experience of valve circuits and whose chief need is to understand, but not to analyse, the information presented.

An experienced reader will understand Mr. Richard's letter, for example, paragraph by paragraph, a less experienced reader will need to move line by line and a child of six or seven will move his finger syllable by syllable. An experienced worker, whether a graduate designer or the service technician, will read the circuit block by block, a timebase, a video amplifier, an i.f. amplifier, etc., the inexperienced worker is analogous to the child in his approach. I believe that there is a need for information to be communicated easily and that the design engineer, being the more flexible, should subordinate his personal inclinations if these conflict with the ready assimilation of information by the non-designer. It must be appreciated that the design staff is a very small percentage of the industrial complex and I view with trepidation the advent of 'transistorised' graduates.

Far from wishing to avoid change, ipso facto, I put forward the view that all progress must be based upon past knowledge and, in a field of valve, p-n-p, n-p-n, and other devices, it serves no useful purpose to the bulk of industry if identical circuits are presented differently according to the nature of the active element. The article "Common Emitter Amplifier" by R. Leek (*Electronic Tech.*, August 1961) serves as a trifling example. The equivalent circuits have elements above the reference line, the actual circuit (Fig. 7) appears to be a cascade of emitter-followers because p-n-p devices are used, if Mr. Leek had used n-p-n devices none of the analysis or equivalent circuits would change but the circuit would now look like a cascade of common-emitter amplifiers. If the drawing office had confused, or omitted, the very small arrow, where should we be?

I have no wish to create 'rules', my reference to the annotation of waveforms merely indicate a route, now discarded, by which a

group of engineers, having considerable valve experience, entered, successfully, into the transistor field. Surely, the use of waveforms within the circuit is normal instructional practice?

My recommendation, not rule, at the moment is to draw the circuit in such a way that the position of the h.t. line is determined by the majority of the active devices in order that the majority of collectors shall be uppermost, that is, negative at the top if p-n-p are in the majority and conversely. There is some slight precedent for this in, for example, a push-pull amplifier where a minority of 'upside-down' devices may be accepted. Additionally, it does conform with present practice as evidenced by the majority of authors in the technical press and in textbooks. I view this as an interim solution until more general agreement can be obtained.

T. G. Clarke.

Radar Development Laboratory,
Decca Radar Ltd.,
Chessington, Surrey.
17th August 1961.

Printed Circuit Plating

Sir,—We were very interested to see Mr. Philpott's article on "Printed Circuit Plating" in your June issue, the more so as we recognise the design on Fig. 1 as one of our own.

We have observed, on continuously-wiped contacts plated with rhodium or palladium, that the contact resistance rises from initial values of tens of milliohms to values of tens of ohms and more, over the first few hundred thousand wipes. The measurement is carried out with an open-circuit voltage of 10 mV d.c. to remove any chance of breaking down any film.

Mr. Philpot does not mention this effect and, of course, for plugs and sockets and manually-operated switches it is of no consequence. We would be interested to know if any of your readers who have noticed this can suggest any theory as to its cause, or any means of preventing it. Gold-plated contacts show no such effect.

M. P. Wax.

Bailey Meters & Controls Ltd.,
Purley Way,
Croydon, Surrey.
1st August 1961.

Sir,—The increase in contact resistance that occurs on continuously-wiped rhodium or platinum-plated contacts, mentioned by Mr. M. P. Wax, is a phenomenon that has interested the writer for some little time.

It has been suggested that this effect is due to the occurrence of a thin film of polymer on the rhodium or platinum surface. Such films are thought to be formed from the organic chemicals that are released by the ambient plastics and that their formation is catalyzed by the platinum group metal.

That gold contacts do not show similar effects, as mentioned by Mr. Wax, is thought to be because gold is far less catalytically active.

J. E. Philpott.

Johnson, Matthey & Co., Ltd.,
London, E.C.1.
16th August 1961.

Magnetic Recording

Sir,—The article 'Magnetic Recording' by Messrs. Sebestyen and Takacs in the August issue of *Electronic Technology* has raised a number of queries in our minds and we feel it might be of general interest if the authors would like to comment on them.

The first query concerns the step between lines (4) and (5) of the mathematical treatment on p. 275. Here it is stated that

$$\tau = \frac{T}{2} + \tau_s \sin \omega t.$$

This implies a movement along the time axis, by an amount $\tau_s \sin \omega t$, of the sloping part of the graph of B_r against t (also on p. 275) which lies between $t = \tau - \alpha$ and $t = \tau + \alpha$. Surely it is truer to say that the application of a signal field will result in a shift of the sloping part between $t = \tau - \alpha$ and $t = \tau + \alpha$ in the B_r direction.

If this shift is given by δB_r then the movement in time of the slope will be given by

$$\delta B_r \Big/ \left(\frac{dB_r}{dt}\right)_\tau \text{ where } \left(\frac{dB_r}{dt}\right)_\tau \text{ is}$$

the gradient of the slope close to $t = \tau$. This gradient will depend on T, the period of the bias oscillation and will be given by

$$\left(\frac{dB_r}{dt}\right)_\tau = \frac{k}{T} \quad \ldots \quad (1)$$

(where k is a constant)

if it can be assumed that a change in T results in a regular compression or expansion of the B_r waveform in time.

If the cyclic swing of δB_r due to the signal field is given by

$$\delta B_r = B_s \sin \omega t \quad \ldots \quad (2)$$

then τ_s can be identified as a term

$$\frac{T B_s}{k}.$$

The signal-voltage output is later given by

$$e = C_1 \omega \frac{\tau_s}{T} \cos \omega t$$

and by substitution this becomes

$$e = C_2 \omega B_s \cos \omega t \quad \ldots \quad (3)$$

where $C_2 = C_1/k$

Thus the dependence of signal output on bias frequency disappears.

The correctness of this modification is supported by the results of the authors' first experiment which indicates that for frequencies of bias between 80 and 200 kc/s the output changes by less than 1 dB per octave of the frequency of the bias. It is perhaps more than a coincidence that the fall of output with a diminishing frequency of the bias appears to start near to that frequency (60 kc/s) at which a point on the tape passing over the record-head gap begins to experience less than one complete cycle of the bias. In this respect it was noticed that, in general, the frequencies used for biasing in the experiments described in the paper were very low by normal standards. This was especially the case in experiment (2) where it would seem that at 30 in./sec frequencies of 5, 10 and 50 kc/s were used.

At the wavelengths associated with these frequencies the authors may be correct in saying that self-demagnetization will be a small effect but, at normal bias frequencies, the wavelength will be shorter by a factor of 4 at least and it would seem almost certain that self-demagnetization would influence recorded waveforms quite considerably.

A further problem is posed by the assertion that the bias field is capable of saturating the tape. If this is the case then a normal record head with bias current flowing in it could be used to erase a modulated bias recording containing signal information. In practice this process does not remove the signal information from the tape although it does produce a few dB reduction in the signal. We feel that it is possible that the authors may have accepted a little too easily the presence of a trapezoidal waveform of induction as indicating with certainty that saturation is taking place.

R. Langhaid.
R. Nichle.

M.S.S. Recording Co. Ltd.,
Caldebrook, Bucks.
17th August 1961.

Transistor Frequency Response

Sir,—Mr. Barry (July, p. 267) has raised some important points concerning our article "Transistor Frequency Response". It must be appreciated that the relative shape of the curves is of most importance when deciding whether a transistor has the greatest potential as a switch or as a small signal amplifier. Because of this, the fact that f_1 does not equal f_T at high collector currents is unimportant.

The contours shown in the paper were similar to those of the 2N1742 (amplifier) and MD501 (switch). The MD501 is normally used to switch currents greater than 6 mA. By extrapolating the contours shown in the article to 20 mA, it can be seen that the switch will give a far better performance in the region 0–2 volts, although the difference in charge required across most of a load line drawn to 20 mA, between the two types will be comparatively small.

The object of this letter is to stress that the importance of f_T contours lies in their shape. We consider this to be a simple, usable way of defining the performance of a transistor at high frequency.

L. J. Bradley.
J. R. James.

Swindon, Wilts.
29th August 1961.

New Books

Colour Television: The N.T.S.C. System—Principles & Practice
By P. S. Carnt, B.Sc.(Eng.), A.C.G.I., A.M.I.E.E. and G. B. Townsend, B.Sc., F.Inst.P., M.I.E.E., A.K.C. Pp. 487 + xii. Published for *Wireless World* by Iliffe Books Ltd., Dorset House, Stamford St., London, S.E.1. Price 85s. (postage 1s. 9d. extra).

"This book describes the British adaptation of the American N.T.S.C. system of colour television. . . . The N.T.S.C. system is explained with particular reference to the 405-line version, but wherever there are differences between the 405, 525, and 625 line systems these are fully explained. . . .

A working knowledge of black-and-white television is assumed, and though the work is largely non-mathematical the more advanced mathematics is given in the appendices. Introductory chapters are included which will enable the reader to understand the principles of colour measurement and the behaviour of the human eye in relation to colour perception. Most aspects of the transmission and reception of colour signals are discussed, though the emphasis is on the latter."

Transistor Circuit Analysis and Design
By Franklin C. Fitchen. Pp. 356 + xi. D. van Nostrand Co. Ltd., 358 Kensington High Street, London, W.14. Price 56s.

This book was designed to be suitable for use as a one semester (six-month) transistor course for engineering students and, as the author writes in his preface, it contains "only that which appears necessary in order to analyse and design the simpler semiconductor circuits".

Chapters 1, 2 and 3 are devoted to the d.c. operating conditions of transistors, and include a very simplified description of the physics of semiconductors which is probably adequate for a book of this nature. Chapters 4, 5, 6 and 7 deal with low-frequency parameters of transistors, and their use in the design of audio-frequency amplifier circuits. Use is made of the equivalent-T circuit, and of the *h*-matrix parameters.

Chapter 8 covers a brief description of simple negative-feedback theory and its application to reducing the spread of gain in transistor amplifiers, and also has a short section on d.c. amplifiers. Chapters 9 and 10 deal with high-frequency applications and introduce the hybrid-π equivalent circuit, and chapter 11 is a short account of the use of transistors in switching and multivibrator circuits.

The book requires no prior knowledge of transistors, but assumes some knowledge of determinants, a grounding in thermionic-valve circuitry, and a working knowledge of a.c. and d.c. circuit theory.

The sections dealing with basic circuits are very clearly written, and the circuit diagrams used for analysis are models of clarity in showing the direction and sign of all currents and voltages. The introductory first chapter specifies the symbols and notation which are used throughout the book, so that there can be no confusion in interpreting the diagrams.

Outside of these sections, the subjects have been so abbreviated that there is very little useful information which would not be known already to readers who have some slight knowledge of valve circuitry. Each chapter ends with a large number of problems, many of which extend the subject matter of the preceding chapter, and a number of references to further books and articles.

There are a number of points at which the text is faulty or inadequate. In the first chapter, in the section 1–6, the author refers to a particular characteristic curve and writes that "it is obvious that when the base current is $-25\ \mu$A, collector current is -1 mA". The current in the characteristic is only too obviously $-1{\cdot}2$ mA.

One point of criticism, concerning the otherwise excellent section on equivalent circuits, is that there is no table of conversion between common-emitter *h*-parameters and common-base *h*-parameters. All the design formulae are expressed in terms of the common-base parameters, and this makes them inconvenient for use when the data given on transistors by a manufacturer is presented only in the common-emitter form.

The section on thermal runaway is most inadequate in dealing with this basic limitation of germanium transistors. The reader is left with the impression that thermal runaway can be prevented by stabilizing the collector current against changes of leakage current with temperature, and the equally important effect of changes of base–emitter voltage with temperature, which can cause thermal runaway even when leakage current is negligible, is ignored. The formula quoted for freedom from thermal runaway (within its limitation of ignoring V_{be}) is incorrect, as the assumption that $\Delta T_j = 0$ is quite invalid under conditions which might produce runaway.

In the section on class B amplification the author states that it is important to know the value of the dissipation at the collector, and then derives a formula which does not give the maximum dissipation but the dissipation under maximum power-output conditions. This is approximately 30% lower than the maximum, and the use of the formula in designing a class B stage could result in exceeding the transistor rating. Also in the same section P_0 represents two different quantities in two successive formulae.

In chapter 10, a diagram of a transistor receiver is described as being of a typical battery-powered a.m. receiver. As two of the transistors are n-p-n, it is certainly not typical of any present commercial design.

The book can be recommended as an introduction to transistor circuitry in the sections which specifically deal with basic circuits, but the reviewer considers the remainder of it is either too sketchy or inadequate even for a short course. W.J.P.

List of Telephone Routes 1961

This list is published by the International Telecommunication Union in accordance with the Telephone Regulations (Geneva, 1958) with an eye to the recommendations issued by the C.C.I.T.T. The List comprises six booklets: List of routes in Europe; Africa; America; Asia; Oceania; and a List of Intercontinental Routes. It appears in a single three-language edition (English, Spanish and French) and the price of one set of six booklets is 17.05 Swiss francs (post free).

Coloured Chart showing the Frequency Allocation between 10 kc/s and 40 Gc/s (Geneva, 1959).

This chart is issued in accordance with No. 813 of the Geneva Radio Regulations (1959). It is very similar in general form to the chart issued by the I.T.U. General Secretariat in accordance with the Atlantic City Radio Regulations (1947). It is published in three editions (English, Spanish and French). Price (per edition) 3.55 Swiss francs.

Both these publications are available from I.T.U. General Secretariat, Palais Wilson, Geneva, Switzerland.

Capacitors, Inductors and Resistors at Radio Frequencies
By T. I. Jones, M.Sc., Ph.D. Pp. 23 + iv. Notes on Applied Science No. 22, published for the National Physical Laboratory by Her Majesty's Stationery Office, York House, Kingsway, London, W.C.2. Price 1s. 9d.

"While most of the sections deal with the properties of resistors, capacitors and inductors some will be devoted to the determination of other quantities by measurement of resistance, capacitance inductance and, on occasions, frequency. It is this which explains why the measurement of the characteristics of radio frequency cables has found a place under the present heading."

Understanding Microwaves (Abridged Edition)
By Victor J. Young, Ph.D. Pp. 292 + xi. Published by John F. Rider Publisher, Inc., New York, and distributed in the U.K. by Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 32s.

" . . . the book is intended to present material which will form a basis for understanding microwave radio and radar . . .

. . . material has been chosen which it is believed will aid those who have not yet considered radio waves shorter than 10 cm and allow them to go on into the various advanced ideas which have been published in great volume and those which are still coming forth."

MEETINGS

I.E.E.

5th October. Presidential inaugural address by G. S. C. Lucas, at 5.30.

6th October. "What is pH and its Measurements?" by Dr. G. Mattock and G. R. Taylor, at 6.0.

10th October. "Trends in Computer Engineering", by W. S. Elliott, at 5.30.

17th October. Discussion on "The Place of Digital Computers in the Teaching of Electrical Engineers" opened by Dr. P. D. Aylett at 6.0.

23rd October. Discussion on "Is Automation Making Satisfactory Progress?" opened by Professor A. Tustin, at 5.30.

25th October. "Global Communication" by R. J. Halsey, at 5.30.

1st November. "Generation of Power in Satellites", by H. J. H. Sketch, at 6 o'clock.

2nd November. "Acoustics and the Electrical Engineer", by T. Somerville, at 6 o'clock.

3rd November. "Information Theory in Relation to Biology." Discussion opened by Professor D. M. Mackay, at 6 o'clock.

These meetings will be held at the Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2.

Brit. I.R.E.

18th October. Symposium on "Digital Differential Analysers", at 6.0 at the London School of Hygiene and Tropical Medicine, Keppel Street, W.C.1.

The Television Society

12th October. "Transistorized Line Timebases for 405- and 625-line Systems", by K. E. Martin.

26th October. "Television from Moscow", by W. Ward and L. F. Mathews.

These meetings will be held at the Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, W.C.2, and will commence at 7.0.

The Radar and Electronics Association

12th October. "Space Communications" Part I, by L. F. Mathews, to be held at 7 o'clock at the Royal Society of Arts, John Adam Street, W.C.2.

Society of Relay Engineers

10th October. "Television Distribution by Coaxial Cable", by G. J. Hunt and C. F. Whitbread, to be held at 2.15 at 21 Bloomsbury Street, W.C.1.

The Institute of Navigation

25th October. Annual General Meeting followed by "Navigation and the Science of the Sea", by Dr. G. E. R. Deacon, to be held at 3 o'clock at the Royal Geographical Society, 1 Kensington Gore, S.W.7.

NUCLEAR ELECTRONICS SYMPOSIUM

The Measurement and Control Section of The Institution of Electrical Engineers is arranging a Symposium on Nuclear Electronics to take place on 30th November and 1st December, 1961. This will be held at the Institution's headquarters at Savoy Place, London, W.C.2, and is intended to follow up the Conference on the same subject held by the International Atomic Energy Agency in Belgrade in May of this year.

Registration forms for the Symposium can be obtained from the Secretary of the I.E.E.

BELLING-LEE 'FILTRON' SUPPRESSORS

Under an agreement concluded with the Filtron Co. of America, Belling & Lee are making a range of suppression filters which have been pioneered and internationally sponsored by Filtron.

These filters are hermetically sealed and suitable for operation in the temperature range 85 °C to −55 °C (some types up to 125 °C). They are made in ratings from 0·15 A up to 300 A and up to 18 kV d.c. working.

Existing Belling-Lee filters will still be available where a lower-performance specification is adequate.

B.S.R.A. SILVER JUBILEE CONVENTION

The British Sound Recording Association has arranged a whole-day audio convention to mark its silver jubilee year. This will be held at the I.E.E. headquarters at Savoy Place, London, W.C.2, commencing at 10.00 a.m. on Saturday, 14th October, 1961. At the Convention five technical papers will be read. Following the Convention the Association's Silver Jubilee Dinner will take place at the Howard Hotel, Norfolk St., London, W.C.2.

Tickets for the Convention and the Dinner can be obtained from Mr. S. W. Stevens-Stratten, F.R.S.A., "Greenways", 40 Fairfield Way, Ewell, Surrey.

S. S. ERIKS, K.B.E.

Mr. S. S. Eriks, Managing Director of Mullard Ltd., has recently been appointed an honorary Knight Commander of the Order of the British Empire.

Mr. Eriks was awarded the O.B.E. in 1948 and was appointed Officer in the Orde van Orange-Nassau by Queen Juliana in 1953.

NON-DESTRUCTIVE TESTING CONFERENCE

A one-day Conference on "New Electronic Techniques in Non-Destructive Testing" is being organized by the West Midlands Section of the British Institution of Radio Engineers. It will be held in the Wolverhampton and Staffordshire College of Technology on Wednesday, 6th December, 1961.

Further information and registration forms may be obtained from the Honorary Secretary of the Section, Major C. W. Weech, M.Brit.I.R.E., 5 Shelton Fields, The Mount, Shrewsbury.

PROFESSOR J. T. ALLANSON

Mr. J. T. Allanson has been appointed to the recently-established second chair in the Electrical Engineering Department in the University of Birmingham. In this post he will be responsible primarily for the undergraduate work of the Department.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations from nominal frequency for August 1961*

Date 1961 August	MSF 60 kc/s 1500 U.T. Parts in 10^9	Droitwich 200 kc/s 1030 U.T. Parts in 10^9
1	− 14·8	+ 4
2	− 14·8	+ 5
3	− 15·0	+ 6
4	− 15·2	+ 5
5	− 15·0	+ 3
6	− 14·9	+ 3
7	− 14·7	+ 3
8	− 14·8	+ 3
9	− 15·1	+ 3
10	− 14·9	+ 2
11	− 14·9	+ 2
12	− 14·7	+ 3
13	− 14·7	+ 2
14	− 15·1	+ 2
15	− 15·1	+ 2
16	− 15·1	+ 2
17	− 14·9	+ 2
18	− 15·1	+ 1
19	− 15·1	+ 1
20	− 15·0	0
21	− 15·3	0
22	− 15·3	− 1
23	− 15·0	− 1
24	− 15·0	− 1
25	− 14·9	− 2
26	− 14·8	− 2
27	− 15·0	− 4
28	− 15·0	− 4
29	− 15·0	− 5
30	− 15·0	− 6
31	− 15·0	− 8

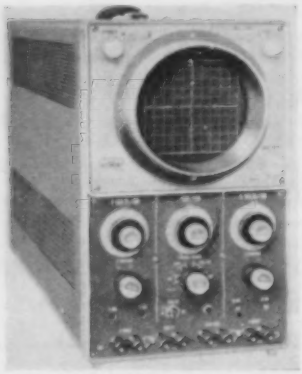
* *Nominal frequency corresponds to a value of 9 192 631 770 c/s for the caesium F,m (4,0) − F,m (3,0) transition at zero field.*

New Products

50 μV/cm Oscilloscope

Aveley Electric have announced the introduction of the DuMont Type 403-B oscilloscope which, it is claimed, is the most sensitive oscilloscope commercially available at present. The maximum sensitivity is 50 μV/cm, with a low noise level which permits the resolution of signals as low as 10 μV.

The Y-channel bandwidth is d.c to 150 kc/s (−3 db±1 db) on the highest-sensitivity range (50 μV/cm) and d.c. to 1 Mc/s on the lowest-sensitivity range (5 mV to 20 V/cm). Its 21 calibrated time-base ranges extend from 1 μsec to 5 sec/cm, with a vernier control providing continuous adjustment between the switched ranges and



also extending the 5 sec/cm range to slower than 250 sec/cm for a 250 sec maximum-sweep duration. A single-sweep mode is also provided. The Y-channel attenuator permits balanced operation on each of its 17 positions; common-mode rejection is better than 120:1.

The X-channel amplifier has a sensitivity of 10 mV/cm at 500 kc/s bandwidth and can be used for phase measurements at frequencies above 100 kc/s.

Other features claimed for the oscilloscope include a nominal drift of less than 500 μV/hr (after warm-up), an a.c. stabilizing control that virtually eliminates d.c. drift, a 'sync ready' indicator for indicating optimum trigger adjustments, a 5 kV c.r.t. with P31 phosphor normally supplied, and a design which does not require selected valves or components.

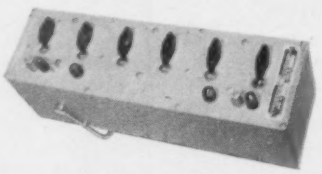
Aveley Electric Ltd.,
Ayron Road, Aveley Industrial Estate,
South Ockendon, Essex.

Ratio Transformers

Two Gertsch ratio transformers which provide an accurate means of supplying and measuring a.c. voltage ratios over a wide frequency range are now available in the U.K. from Wayne Kerr.

Known as the RT4 and RT5, the transformers operate from 30 to 1,000 c/s and 50 to 10,000 c/s respectively. Over the lower frequencies of each range, the ratio accuracies are 0·001%; 0·01% accuracy is obtainable over the higher frequencies of each range.

Ratios are selected by operating five decade switches and a single-revolution potentiometer. This arrangement provides continuous resolution and readout to six significant figures, each ratio value being shown as a direct decimal reading. Both units have a terminal linearity of 0·001% and a very small phase angle under loaded conditions (depending upon the frequency and ratio settings).



The effective series impedance of the RT4 is 4 Ω and its input inductance is 500 H; comparable figures for the RT5 are 3 Ω and 150 H. The normal temperature range for both transformers is from 80 °C to −15 °C.

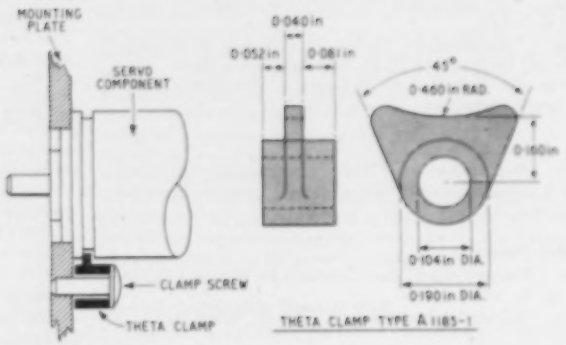
Wayne Kerr Laboratories Ltd.,
44 Combe Road,
New Malden, Surrey.

Servo Mounting Clamps

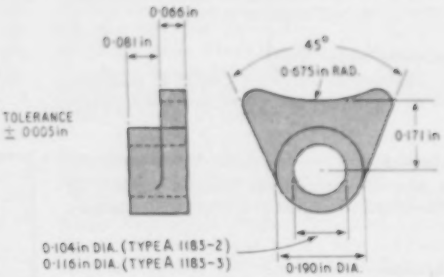
Two standard mounting clamps for synchros, servomotors and potentiometers are now being marketed by the Theta Instrument Corporation, U.S.A.

These clamps cover housing sizes 8, 10, 11 (commercial and military), 15 and 18. Each unit is machined from solid aluminium stock.

Theta Instrument Corporation,
520 Victor Street,
Saddle Brook, N.J., U.S.A.



THETA CLAMP TYPE A 1183-1



THETA CLAMP TYPE A 1183-2 & TYPE A 1183-3

High Sensitivity On/Off Controller

A controller which is based on the interruption of light falling on to a phototransistor by a small flag attached to a normal meter movement pointer is being produced by Measuring Instruments (Pullin) Ltd.

The unit is designed to provide on/off control of any occurrence which can be measured by a suitable transducer having an electrical output which can be fed to the controller; the maximum sensitivity of the controller is 20 μA.

The controller has two output circuit arrangements: the first is an isolated single-pole double-throw switch rated at 250 V, 3 A a.c., and the second is a circuit switching a mains supply of 200–250 V a.c. at 3 A.

In operation the meter and the associated control relays etc., can be mounted together in one box or alternatively the meter can be remote from the control box connected by a 6 ft. multi-conductor cable.

Three basic models of the controller are available: one with the sensing unit at the maximum end of the meter scale, the second

with the sensing unit at the minimum end of the scale or the third model with sensing units at both ends of the scale. In addition moving-coil or moving-iron movements of various sensitivities are also available.

Measuring Instruments (Pullin) Ltd.,
Electric Works,
Winchester St., London, W.9.

Decade Resistance Box

A decade resistance box which has been designed primarily as a low-cost unit of reasonable accuracy has been introduced by Lionmount for d.c. and audio-frequency applications.

This is the type RD2, which is a 4-decade switched unit covering the range 1 Ω to 11,110 Ω with an accuracy of better than 0·2% (at 20 °C) on the three highest decades

and 1% on the lowest decade. The maximum permissible current rating is 20 mA for the highest decade, increasing to 600 mA for the lowest decade. The unit uses switches having a minimum contact resistance of less than 0·006 Ω and is constructed in a robust aluminium-lined case measuring 12 in. × 3 in. × 3¼in.

Lionmount & Co. Ltd.,
40 Hatton Garden,
London, E.C.1.

Pulse-Measuring Potentiometer

For the accurate measuring of high-voltage pulses in radar transmitters and other pulse circuitry Ferranti have introduced the type PD.30 Pulse Measuring Potentiometer.

This instrument consists of a high-voltage vacuum capacitor, internally shielded, mounted on top of a second element which can be either a capacitor for high-voltage division or a resistor for differentiation.

The PD.30 is designed to handle voltage pulses up to 30 kV and has four switched ranges giving two pre-set voltage division ratios and two differentiation ratios; voltage ratios are normally pre-set at 250:1 and 500:1 and differentiation ratios at 7·5 kV/μsec/V and 5 kV/μsec/V. In the design, allowance has been made for the shunting effect of the input capacitance of the associated oscilloscope so that on the most sensitive range a change of 8·5 pF produces approximately a 1% error. The fixed capacitor has a value of 3·4 pF ± 0·2 pF and the total load capacitance is 6 pF maximum.

The potentiometer can be supplied in a form suitable for mounting into equipment or in a case.

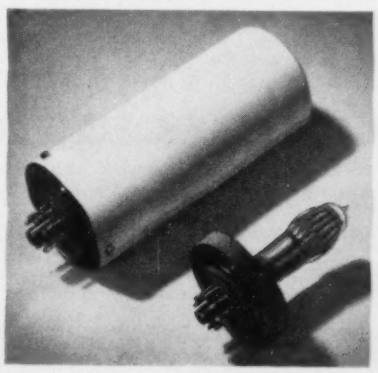
Ferranti Ltd.,
Hollinwood, Lancs.

Crystal Oven

An octal-based crystal oven which has a cyclic temperature variation of ± 0·0014°C is now available from Marconi's Wireless Telegraph Co.

Known as the crystal oven Type F3006, it has an improvement factor of not less than 250:1 (improvement factor is the ratio of the change in outside temperature to that occurring within the oven). The operating temperature of the oven is 79·5°C and the time taken to come within 0·2 °C of this is 35 min. The switching cycle at 20 °C ambient is 20 sec on and 80 sec off; the heater power supply is 24–30 V 50 c/s with a peak consumption of 30 W.

The performance of the F3006 oven is achieved by making use of the properties of a substance (in this case naphthalene) which changes from the solid state to the liquid at a particular and constant temperature. While the melting point of naphthalene is always a constant, the volume of the substance increases in the transition from solid to liquid. It is this increase in volume that is used to control the applied heat so that in practice the temperature is maintained at the melting point and is therefore stable.



Marconi's Wireless Telegraph Co. Ltd.,
Chelmsford, Essex.

Miniature Moving-Coil Relay

Leland Instruments have announced a Microtact hermetically-sealed miniature moving-coil measuring relay which is available with three alternative windings to provide full scale deflection sensitivities of 50 μA, 250 μA and 2·5 mA.

The relay comprises a moving-coil galvanometer with an adjustable contact fitted to the moving coil and fixed contacts

at either end of the arc of travel. The whole assembly is enclosed in a cylindrical metal container which is mounted on a standard 7-pin miniature valve base; measuring 40 mm long × 20 mm diam. the relay weighs 1·12 oz.

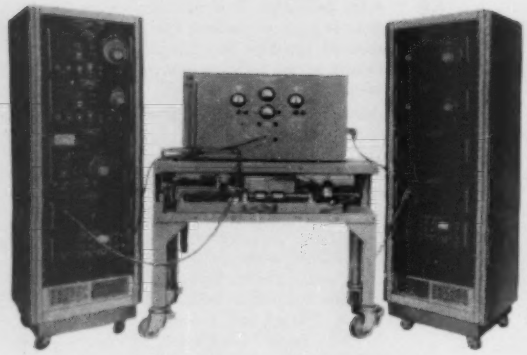
In addition to the three basic models, units can be supplied with the adjustable moving contact pre-set, at the time of manufacture, between the minimum fixed contact and the maximum contact; in the most sensitive model the adjustable moving contact can be advanced to a point where only 5 μA is required to close the moving–max. fixed contacts.

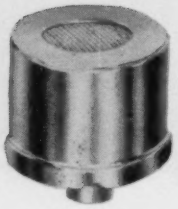
Leland Instruments Ltd.,
145 Grosvenor Road,
London, S.W.1.

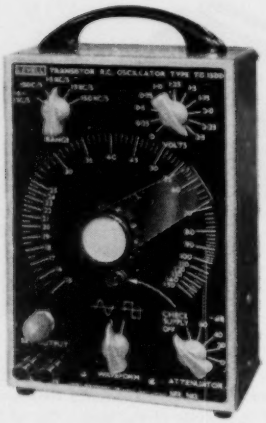
U.H.F. Attenuation-Measurement System

An accurate easy-to-operate system for measuring attenuation in the u.h.f. region has been developed by Weinschel Engineering, U.S.A. This system, the VM-1 Standard Attenuator Comparator, can be used over the frequency range 100 Mc/s to 12·4 kMc/s by selection of suitable mixers, signal sources and oscillators (all of which are produced by Weinschel). It covers directly in one step the voltage range of 40 mV to 0·2 μV; corresponding to power levels of −15 dBm to −121 dBm in a 50 Ω system.

In operation, a local oscillator is used for heterodyning the signal being examined, leaving a resulting 30 Mc/s signal. A separate 30 Mc/s reference signal is fed through an accurate piston attenuator to a comparison







head. This attenuated reference signal is compared with the 30 Mc/s output of the mixer and the change of attenuation required in the piston attenuator to adjust the system to a null is an accurate direct measurement of the r.f. level change.

The VM-1 is direct-reading in decibels and needs no charts or calibration curves. The frequency drift of the r.f. source is compensated for by automatic frequency control of the local oscillator.

Typical setting-up time for one frequency is 15 min. after which, a normally-skilled operator can calibrate eight 10 dB steps of a signal-generator output attenuator in about 3 min.

Weinschel Engineering Company,
Kensington, Maryland, U.S.A.

Dry-Reed Relay

A dry-reed relay which has been produced for printed-circuit applications has been announced by B. & R. Relays.

This is type R.03, which is mounted on a nylon bobbin and measures 1 in. diam. and 2 in. long (excluding reed ends). The relay contains one normally-open reed with contacts rated at 15 W (d.c. resistive), the maximum current rating is 1 A (15 V d.c.) and the maximum voltage rating is 250 V d.c. (60 mA). Coils may be specified to

requirements with resistances up to 20 kΩ and, while the nominal power of 0·06 W is required, the relay will operate efficiently at 0·048 W. The life expectancy is 20×10^6 operations at maximum contact rating and a contact resistance of 0·03 Ω.

B. & R. Relays Ltd.
Temple Fields, Harlow, Essex.

Ultrasonic Transducer

A small inexpensive 40 kc/s ultrasonic transducer is now being produced by Cohu Electronics, U.S.A. This is the model TR-7 which, as a transmitter, produces a sound-pressure field at about 25 dB above 1 μbar at 1 ft with 25 mW excitation. The unit's receiving sensitivity is approximately 2 mV/μbar. The continuous power rating for the unit, with approximately 50 V r.m.s. applied, is ½ W; the intermittent power rating, with a 10% duty cycle and 150 V r.m.s. applied, is 5 W.

With a tuning choke connected, the unit has a bandwidth of about 4 kc/s. It is 1 in. long × 1 in. diameter and weighs ½ oz.

Massa Division of Cohu Electronics Inc.,
280 Lincoln Street,
Hingham, Mass., U.S.A.

Transistor Clips

A range of mounting clips for various types of transistor is now being produced by Jermyn Industries.

These clips are made of beryllium copper and are silver-plated. They include models with diameter ranges from 0·189 in. to 0·2 in. (OC44 to OC201) and length of ½ in. or ¼ in.; a model covering 0·305 in. to 0·358 in. diameter for the TO5 Jedec case, also the OC170; and one to accommodate the TO18 Jedec case. Certain of these models

are also available fitted with an anti-rotation lug and in different gauges of beryllium copper.

Jermyn Industries,
74 Lakeside Drive,
Bromley, Kent.

Sine and Square-Wave Oscillators

Two versions of a combined sine and square-wave transistorized oscillator which covers the frequency range 1·5 c/s to 150 kc/s have been announced by Levell Electronics.

They are known as types TG150DM and TG150D and they are the same except for the inclusion of an output meter in the type TG150DM.

The square-wave signal amplitude, which is directly proportional to the supply voltage, is continuously variable up to 2·5 V peak at a source impedance which varies from zero to 600 Ω at maximum output.

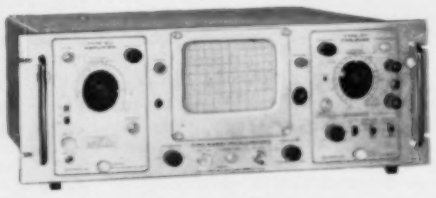
When being used as a sine-wave source, the output is continuously variable up to 2·5 V into 600 Ω and the accuracy is ±(3% + 0·15 c/s) at 25 °C.

Both versions measure 10 in. × 6 in. × 4 in. approximately and are provided with two type PP9 batteries which supply the required power (18 V at 12 mA) for about 400 hr normal use.

Levell Electronics Ltd.,
10–12 St. Albans Road,
Barnet, Herts.

Rack-Mounted Oscilloscope

A rack-mounted oscilloscope which has been designed to accept a wide range of plug-in units in both channels has been announced by Tektronix.



The basic indicator unit, Type RM561, includes: a 5 in. rectangular c.r.t. with a viewing area of 8 × 10 cm and operating with an e.h.t. of 3·5 kV; *z*-axis input; and amplitude calibrator; and regulated d.c. heater and h.t. supplies.

The indicator accepts any of eight available plug-in units. These include a basic amplifier, 1-Mc/s amplifier, differential-input amplifier, dual-trace amplifier, wide-band amplifier and two timebase units.

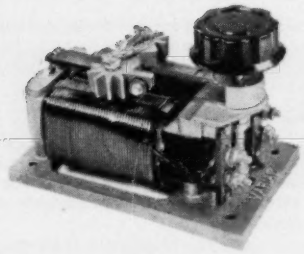
The dimensions are 7 in. × 19 in. × 18 in. and the unit is designed to operate from mains supplies of 105 to 125 V or 210 to 250 V, 50 to 800 c/s and to provide 90 W of power for the plug-in units.

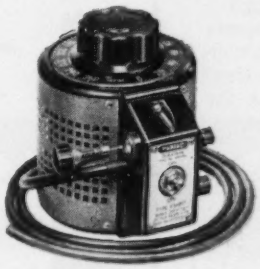
Tektronix, Inc.,
P.O. Box 500,
Beaverton, Oregon, U.S.A.

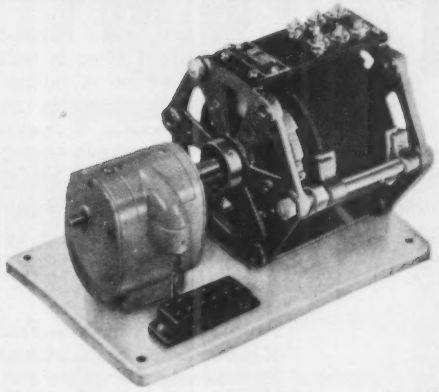
AN IMPORTANT ANNOUNCEMENT

TO USERS OF Variac

THE ORIGINAL AND BEST VARIABLE TRANSFORMER







'Variac' (Regd.) continuously-adjustable auto-transformers have for many years been available from two sources, The Zenith Electric Co. Ltd., of London and Claude Lyons Ltd. of Liverpool and Hoddesdon.

An arrangement has been made whereby, from the 1st October 1961, all manufacture of 'Variacs' in the United Kingdom will be carried out by The Zenith Electric Co. Ltd., and all sales and distribution by Claude Lyons Ltd.

This new arrangement will benefit all customers of both companies, by increased manufacturing and distributing efficiency, and by making available to all the combined technical knowledge and experience of both companies.

As from the 1st October, 1961 enquiries and orders for Zenith 'Variac' transformers, from customers in the United Kingdom, Eire, and British Colonies should be addressed to the exclusive distributors:—

Claude Lyons Ltd.

76 OLD HALL STREET · LIVERPOOL, 3
TEL: CENTRAL 4641-2 TELEX 62181
VALLEY WORKS · HODDESDON · HERTS
TEL: HODDESDON 4541 (6 LINES) TELEX: 22724

(export enquiries to the manufacturers: The Zenith Electric Co. Ltd., Villiers Road, London, N.W.2. Tel: WILlesden 6581-5)

IMPORTANT

This is an **arrangement for the marketing of Variac Transformers only.** There is no policy change in regard to any other product of either of the two companies, and all their other activities will continue independently.

. . . a radar wave directed through a complex layout, often in restricted space, call in 'Waveflex'—a unique advance in waveguide technique. In the 'Waveflex' range of Flexible Waveguides are the answers to a hundred-and-one problems of complicated waveguide plumbing.

The three types of WAVEFLEX Flexible Waveguides—

Twistable—flexible in E and H planes, and twistable about the longitudinal axis.
Non-twistable—flexible in E and H planes.
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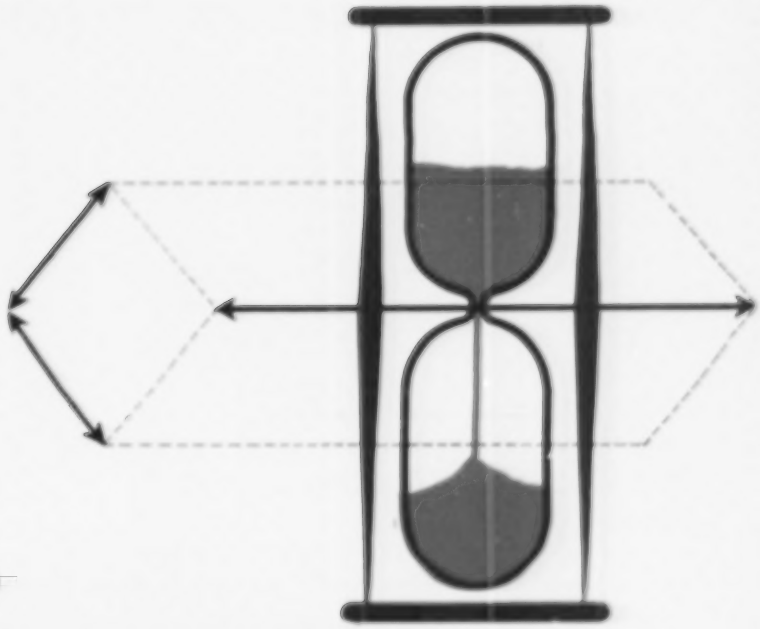

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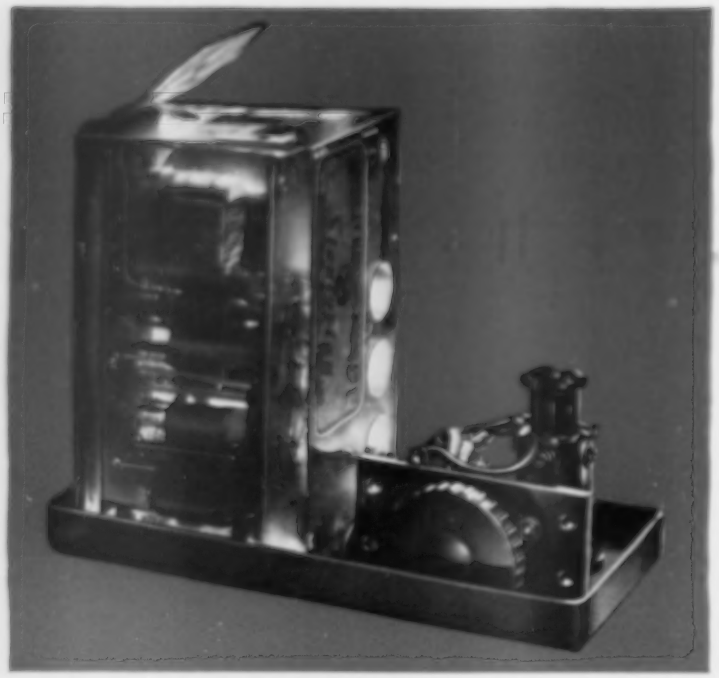

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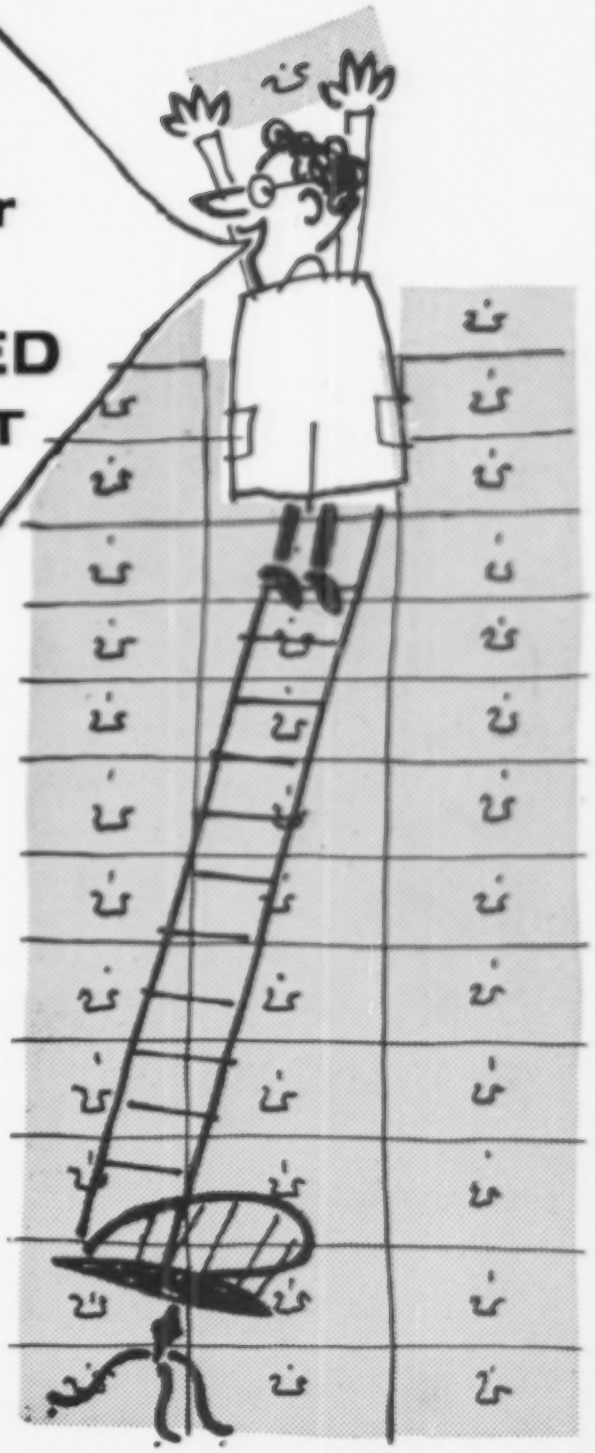

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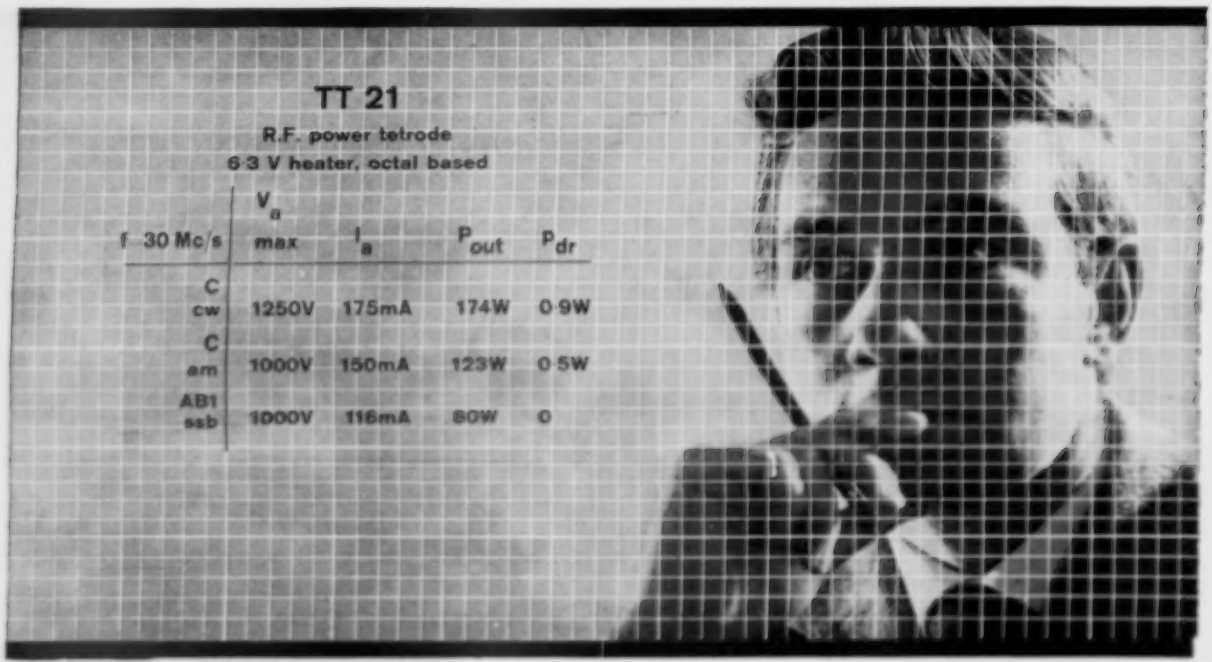

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